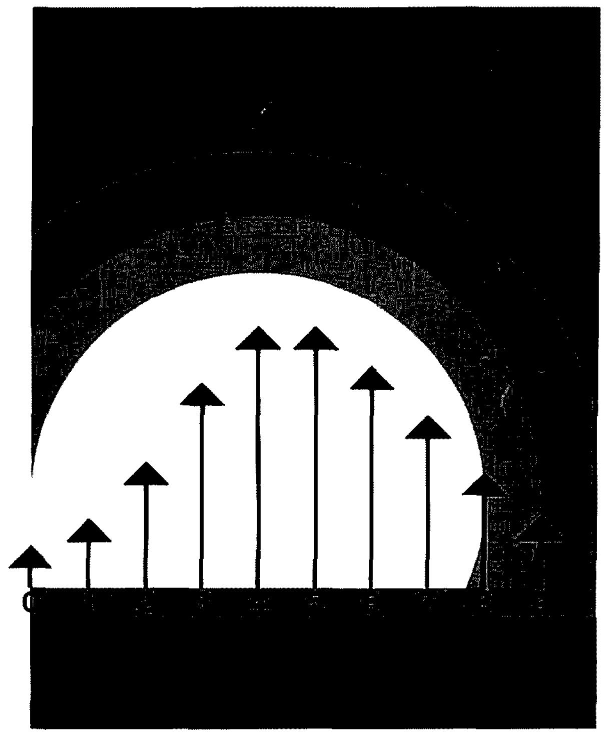
one dollar

# ham radio magazine

## JANUARY 1977





# an introduction to single-sideband fm

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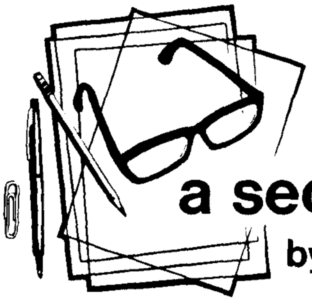

# a second look

by Jim Fisk

In the evening when I'm working down in the shop, I usually turn on the stereo and listen to some nice, easy music. When I get tired of that, I flick on the receiver and tune around the upper end of 75 meters for an interesting round table – I can invariably find an interesting conversation, whether it's about the sorry state of the economy, the high cost of fuel oil, the fine golfing weather in south Florida, or a technical discussion on the merits of quads vs Yagis. Whatever my mood, there's always something on 75 that piques my curiousity, and occasionally something that starts the old adrenalin flowing.

The other night, for instance, two characters with two-letter calls were holding forth on the low end of 75 phone just inside the Extra class segment. The technical topic of the evening was transmission lines, and to hear these guys talk, they were the *original* experts. In reality all they had to offer was a barge full of baloney. Now I'll readily admit that, of all the subjects in amateur radio, transmission lines are among the most difficult to understand, but that isn't a license to run off at the mouth. And to be perfectly blunt, some amateur publications suffer from the same foot-in-mouth disease when it comes to transmission lines and antennas.

I don't know where all the feedline myths started, but I suspect it had something to do with the do-it-yourself swr bridges which first became popular back in the early 1950s. Up until then most amateurs didn't even know about standing waves, and if they did, they didn't seem to care. However, swr bridges soon caught on, and it wasn't too long before being caught with your swr up was synonymous with getting caught with your pants down. At one time someone even suggested that the Q-signal QSW be used with a scale of 1 to 10 to report your latest swr reading as you moved around the band. Fortunately the idea never caught on.

Some amateurs got interested enough in the subject of standing-wave ratios to dig into the books, but when they discovered that swr is a result of power reflected from a mismatched antenna, it only served to reinforce the myth. If a mismatched antenna causes power to be reflected back down the line, they reasoned, this power *obviously* wasn't radiated by the antenna. Some even suggested that the reflected power got back into the transmitter tank circuit and was dissipated in heat; others apparently thought that reflected power was lost forever to some great swr heaven in the sky. A few well-informed amateurs tried to nip these absurdities in the bud, but it was hopeless – the disease spread faster than the cure.

The whole subject of transmission lines is much too complex to be covered in this short space, but it's time to bury some of the myths. First of all, reflected power is not lost nor does it heat up the tank circuit of your transmitter. Secondly, if your feedline has low loss as is the case on the hf bands, increased loss due to swr is so small you can forget about it. Since a 10:1 swr on 100 feet of RG-8/U at 4.0 MHz increases loss by *less than 1 dB*, don't worry about the fact that the swr rises above 2:1 at the band edges – the station at the other end won't be able to tell the difference. If your transmitter doesn't like to load into a mismatch greater than 2:1, buy or build an antenna tuner and save yourself a lot of grief by forgetting the swr on the line to the antenna if it's within reasonable limits, say 10:1. And finally, if you don't understand transmission lines, don't make wild statements before you have your facts straight.

Jim Fisk, W1DTY
editor-in-chief


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More Details? CHECK—OFF Page 126

# presstop

FCC CONSIDERS BAN ON LINEAR AMPLIFIERS for all services, including Amateur. FCC Chief Engineer Ray Spence dropped that bombshell in response to a question posed at the Radio Club of America annual banquet's "Round Table 1976 — Personal Communications" session. At the New York meeting, Ray said the continued circumvention of the FCC's ban on "CB" linears by half a dozen or so manufacturers is forcing the Commission to take a much stronger stance which could result in an across-the-board prohibition on amplifiers for everyone.

Alternatives To Be Considered in the proposed rule making include, in addition to the possible ban, putting all Amateur Radio equipment of commercial manufacture under the equipment approval program that presently applies to gear used by other services. Still another approach would be to forbid the sale of an amplifier to anyone not holding a license that authorizes his use of that amplifier, and make the dealer liable for any sales made in violation of that prohibition.

An Industry Group Dedicated to the well-being of linear amplifiers has been formed. Goals of the new group, being spear-headed by Dentron, are to establish "industry regulations" for the proper use of amplifiers, develop marketing techniques to keep them out of the hands of illegal users, and work closely with the FCC to assure enforcement of the Rules regarding linears.

1X2 CALLSIGN APPLICANTS newly eligible January 1 can be assisted by another ARRL prepared chart showing callsign availability. The chart will be available late in the month — send an SASE marked "1X2 Callsign Matrix" to the League for a copy.

1190 1X2 Callsigns have now been granted since July 2, with 227 applications still pending. No NX2s have yet been issued as the computer program is still being worked out, though applications for the new prefix have been received.

FCC REDUCED the number of copies it requires to be filed in Commission rulemaking procedures from 12 to 6 (an original and 5 copies) in an action announced November 9. Anyone wishing a copy of his comments to go to each Commissioner individually may still file an original and 11 copies as before, but the 6 copies specified in the new ruling assures that a filing is distributed to the appropriate FCC staff members.

The Relaxation, which amended Part 1 of the Rules, became effective November 22.

PORTABLE AND MOBILE DESIGNATION use ended officially on Friday, November 26, a Thanksgiving present from the FCC, but you'll still need to have an "active" mailing address so you can answer possible FCC citations or advisories within the required 10-day period. Also, portable and mobile designations will still be required of participants in ARRL-sponsored contests. Though portable and mobile designations are more important to some contests than others, it's likely the Rule will be across the board for League events.

NOVICE LICENSE TURNAROUND time can be sharply reduced, say several sources, by printing "AMATEUR RADIO NOVICE APPLICATION" in large letters on the envelope used to apply for a Novice exam, then returning the completed exam to Box 1120 in the envelope it came in marked "COMPLETED AMATEUR RADIO NOVICE EXAM ENCLOSED."

NOVICE CLASS INSTRUCTORS CAN REQUEST written examinations for their students before the students take the code test under a Rules waiver announced by the FCC. Under the terms of the waiver, which will be in effect until June 30, an instructor with a class of five or more students should request sufficient exams for his students at least 30 days before the date he plans to administer the exam. Present FCC requirements for examiners remain unchanged — 21 or more years old, a General or higher class license — and the examiner will be held responsible for returning unused exams unopened along with the exams that have been taken by the students.

Requests For Group Exams should include a photocopy of the examiner's license along with his name and address, the number of exams required, and the date the exams are to be given. To facilitate processing by the FCC, the ARRL recommends including two mailing labels which include the above information along with each request. Requests should go to the FCC, Box 1020, Gettysburg, Pennsylvania 17325.

15TH ANNIVERSARY OF AMATEUR RADIO IN SPACE occurred in December — OSCAR 1 was launched December 12, 1961. OSCAR 7 has now provided more operating time for users than all previous OSCARs combined, according to W3PK (K3JTE).

Tests Of JAMSAT's 145-435 MHz transponder have been going very well, and the Japanese-built unit has met specs nicely.

ITOS-1 Launch, which AMSAT had expected to provide this summer's OSCAR launch, has been scrubbed. LANDSAT C looks like the probable alternative.

# an introduction to single-sideband fm

Single-sideband fm, though still in its early stages of development, may prove useful for radio communications — here are the basics of the ssb fm system

**During the past decade** a new communications system was mathematically shown to be possible, was analyzed in depth, and finally several working systems were built in the laboratory. This system, which will be described in some detail in this article, has come to be known as single-sideband fm or (or ssb fm), a term which is also used to describe single-sideband phase modulation.

It is interesting to note that, even though laboratory ssb fm generators exhibit unwanted sideband suppression of 35 dB or more and are relatively straightforward to build, no one ever actually set out to use ssb fm in on-the-air trials. In fact, most professional interest in ssb fm seemed to disappear by 1971 or 1972, perhaps because it was being viewed primarily as a means of saving spectrum space in the commercial fm broadcast band.

Before discussing such details any further, it is necessary to describe the basic system, and to give some historical background as well. Considering the current interest in amateur vhf, fm, the subject of ssb fm should be of general interest, and it is hoped that the following discussion will lead to some amateur experimentation with this mode of communications.

It is not really clear who invented ssb fm, although many people credit it to K.H. Powers, whose work at RCA led to a patent application which was filed in March, 1958.[1] The patent was not granted until September, 1962 – a happenstance which muddied the waters a bit, because E. Bedrosian of the RAND Corporation almost simultaneously published similar results in the *Proceedings of the IRE*.[2] Since Powers' results were never published, Bedrosian was generally credited with the invention. Certainly, Bedrosian's impressive contributions to the theory of signal processing were extremely important (ssb fm was a "spin-off" of his theory, as was an analysis of clipped waveforms in ssb a-m) and it is probably safe to say that it was his work which most directly influenced the subsequent development of the system.

In 1964 Dubois and Aagaard published an article which further analyzed ssb fm and which also contained block and schematic diagrams of a working system, along with appropriate oscilloscope patterns.[3] Although they were only operating at 500 kHz, a practical ssb fm generator had clearly been developed! Less than a year later, Glorioso and Brazeal of the University of Connecticut published further details and discussed the performance of an ssb fm generator they had built.[4] They were among the first to point out the need for a properly designed ssb fm receiver.

The question of how best to receive ssb fm was also getting attention at Princeton University, and during 1966 Kahn and Thomas published an article in which the theoretically optimum ssb fm receiver was discussed.[5] Although the theoretical receiver was impossible to build (as was pointed out in the article), it served as a guide and showed that a modified PLL detection system was the correct route to take.

Hence, by the end of 1966 most of the groundbreaking had been completed, and it seemed inevitable that ssb fm would become a topic of keen interest to professional and hobbyist alike. Paradoxically, by the end of 1971 the system had virtually faded from further discussion in the technical journals, and a search of the

By Richard R. Slater, W3EJD, RD 3, Cherry Lane, Export, Pennsylvania 15632

standard library reference indices shows that no general-interest electronics or amateur magazine ever published any material on ssb fm.

## technical discussion

Throughout the remainder of this article, I am going to assume that you are at least partially familiar with the sideband representation of ordinary fm. (If you wish to brush-up on the properties of fm sidebands, see any recent *Radio Amateur's Handbook*). Also, I intend to avoid the use of any mathematics, even though that means some loss of accuracy in the discussion, and will use as examples only sinusoidally modulated signals with full carrier. The full mathematical details can be obtained by consulting the list of references.

## the ssb fm signal

Although most people may be able to guess what the spectrum of an ssb fm signal ought to look like, a true starting point lies in defining the way in which it must behave. From the journal articles already mentioned, it turns out that a true ssb fm signal has two important properties:

**1.** It must be possible to detect the signal with a device which is sensitive only to frequency deviations (for example, an ideal ratio detector).

**2.** The transmitted signal must contain only upper or lower sidebands, but not both.

These requirements are not particularly surprising until some thought is given to the second one. If we can really find a way to eliminate one group of sidebands, then it

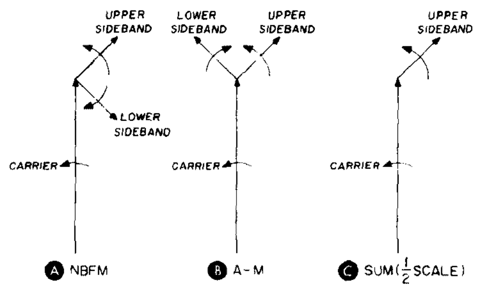


**fig. 1. Rotating vector for narrow-band fm (nbfm) and amplitude modulated (a-m) wave at a modulation index of 0.4 and at 40% modulation, respectively. The same oscillator is assumed to produce both carriers, so that the carrier vectors remain in phase.**

will no longer be true that, at each instant of time (as in conventional fm), all sidebands add together to give a constant-amplitude signal. Stated another way, a ssb fm signal must have both its amplitude and its frequency change with time. To most people this is the first surprise. The second surprise comes when the first requirement *is then re-examined*: It implies that, if we run the ssb fm signal through a limiter (and thereby remove the amplitude changes), the signal is converted to ordinary fm. If this were not so, then the detector demanded by the first requirement wouldn't be capable of recovering the modulation.

The amplitude variations in ssb fm can perhaps be made a little more palatable by recalling that a ssb a-m signal also changes in frequency *and* amplitude. In other words, sideband signals of any type are pretty uncooperative waveforms: They simply refuse to retain the familiar characteristics of their parent signals, especially when viewed on an oscilloscope.

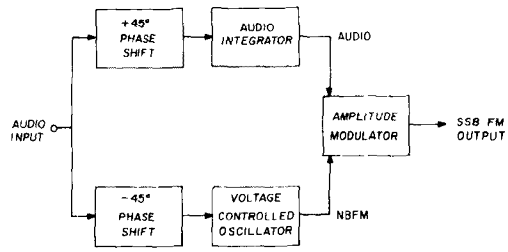


**fig. 2. Block diagram of a suggested ssb fm generator (see text).**

The ability of a *limiter* to convert ssb fm to ordinary fm will hopefully seem perfectly reasonable after the actual method of signal generation has been discussed.

## generation of ssb fm

Most of us have, at one time or another, seen the rotating vector model for an a-m waveform. This was widely used in the past to explain the operation of the phasing method of ssb a-m generation; it also is a useful way of introducing ssb fm as well because the methods for generating ssb fm introduced to date are all based on phasing systems.* Since fm signals can contain several hundred sidebands when modulated with a single tone, it is necessary that we not consider the general case right away. Instead, to simplify the concept of rotating vectors, consider the problem of "sidebanding" a true nbfm signal. This, by mathematical definition, is an fm signal which has only one significant pair of sidebands present; the corresponding rotating vectors are shown in **fig. 1A.** These vectors are identical to those often used in descriptions of narrow-band phase modulation because they can represent either case.

***Fig.* 1B** shows a diagram for an ordinary a-m wave simultaneously modulated by the same audio tone. The implication is obvious: If all of these vectors are added together, the lower sideband will cancel, leaving an ssb signal with carrier. The next problem is to decide which kind of ssb it is, since it is the result of both an nbfm and an a-m wave.

The question can be answered by asking ourselves what would happen if it were run through an imaginary limiter/fm detector. Clearly, the amplitude variations would be removed and only the frequency excursions would remain. The frequency excursions which are still

*For reasons too involved to be explained in this introductory article, filtering methods are not satisfactory.

present occur at the same rate as the original audio so we conclude it is ssb fm.

But what if the same signal is run through an a-m detector? The signal will have only its amplitude variations recovered, but these too, are present and occur at the audio rate. Thus, we must also have ssb a-m!

The point is not made to confuse (though that may be its present effect), but rather to show that, when only one sideband remains, the presence of a properly phased carrier can enable either an fm or an a-m detector to recover the audio.* We are therefore at a departure point and must introduce multiple sidebands if we are to obtain a sideband signal which is distinctively different from ssb a-m. However, the method is basically correct and, for the moment, we can use the vector diagram to investigate some of the problems of actually using simultaneous a-m and fm to produce a single-sideband signal.

The very first problem which arises lies in the non-linear nature of fm: If everything else is held constant, and only the frequency of the modulating tone is changed, the amplitude of the fm sidebands will still change (recall that, in nbfm, the sideband amplitude is inversely proportional to the modulating frequency). Thus, if the lower-sideband of **fig. 1** has been carefully nulled out at a particular modulating frequency, it will reappear whenever that frequency is changed.

It is obvious that some means is needed to make the system frequency-independent. One method which can be used is to modify the a-m section so that the a-m

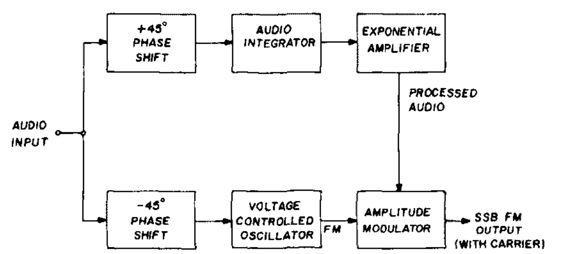


**fig. 3. System block diagram of a working ssb fm generator (after Dubois and Aagaard, reference 3).**

sidebands will also change in amplitude as the modulating frequency is changed. If this can be done in just the right way, the a-m lower-sideband will exactly cancel the fm lower-sideband at any audio frequency.

In the case of frequency modulation, it is known that the modulation index decreases by one-half each time the modulating frequency is doubled. In the narrow-band case, the sideband amplitudes are proportional to the modulation index if a constant deviation is maintained. Hence, the nbfm sidebands decrease in amplitude at 6 dB per octave as the modulating frequency is increased. If the a-m sidebands can be made to behave in the same way, then the desired result will be obtained. In principle, this is easily accomplished by inserting a simple R-C lowpass filter into the a-m audio section. However, there is a very real problem which arises whenever an R-C lowpass filter is used in this way – it will be a source of audio phase shift. This phase shift will rotate the a-m vectors of **fig. 1** so that they no longer cancel the unwanted sideband. Furthermore, the audio phase shift may depend upon the modulating frequency.

In order to control phase shift, a particular kind of lowpass filter – known as an *integrator* – may be used. This introduces a constant phase shift of about 90 degrees at all audio frequencies of interest. The problem of phase matching can then be solved by the introduction of another 90-degree, all-pass network somewhere else in the system. An ordinary R-C filter can be used as an integrator provided that its time constant is at least 15 times the period of the lowest frequency component in the audio. However, the very high attenuation which results can be a disadvantage.

Unfortunately, the 90-degree all-pass network cannot be realized in practice, so commercially available audio phase-shift networks are used which produce two outputs of plus and minus 45 degrees with respect to the input audio. **Fig. 2** is a block diagram of a suggested system.

## a working system

Throughout the discussion to this point, it has been incorrectly assumed that only two sidebands have to be considered in the case of nbfm. While it is true that only two sidebands are present as long as the modulation index is smaller than about 0.4, it is *not* true that what an amateur calls nbfm always has such a small modulation index.

Recall again that, every time the modulating frequency is halved, the modulation index doubles. This means that speech – which contains many audio frequencies – will probably contain a few frequencies which are low enough that the modulation index will considerably exceed 0.4 and which will, in turn, produce multiple sideband pairs. In such a situation, the a-m section won't have a "matching" sideband pair available, and cancellation of all unwanted sidebands will be impossible. In fact, with the fm systems used by most amateurs, nearly every audio frequency produces more than one pair of sidebands. Clearly, the hypothetical system of **fig. 2** will not do the job, and something additional is needed.

## the exponential amplifier

It was at this point that Bedrosian's theory made one of its important contributions; If the audio applied to the a-m section is intentionally distorted, you can generate multiple a-m sidebands with frequencies which match those of their fm counterparts (the distortion generates audio harmonics, which in turn generate the required additional a-m sidebands). The distorting of the audio

*Lest an innocent reader attempt to detect *suppressed-carrier* ssb a-m on his fm receiver, let me hasten to add that an injected carrier is unable to maintain the proper absolute phase for fm detection and further, it must be injected before limiting takes place.

signal has to be done very precisely, and the correct way of doing it is not very obvious. It is a direct result of Bedrosian's theory that the proper method of doing this is to use a so-called exponential amplifier in the audio chain of the a-m section. The exponential amplifier is not difficult to build, and it turns out to be a kind of inverse compression amplifier. It operates by increasing its gain as the audio input voltage increases. In fact, the gain should change as the mathematical exponent of its input (hence its name). Luckily, this rather exotic-sounding behavior is obtainable simply by inserting a properly-biased diode in a feedback loop.

Fig. 3 shows the corrected system, and this arrangement does produce ssb fm. It is basically the same arrangement used by Dubois and Aagaard in their pioneering experiments in the middle 1960s.[3]

Snider and Schilling[6] have published an IC version of the exponential amplifier (which they called an *exponentiator*); a schematic diagram of their circuit is shown in fig. 4 to illustrate the ease of construction (the three transistors are an amplitude modulator). Note that the 1N205 diode must have a 150 millivolt drop across it to provide correct operation, and that the last IC must be dc-coupled to the amplitude modulator.

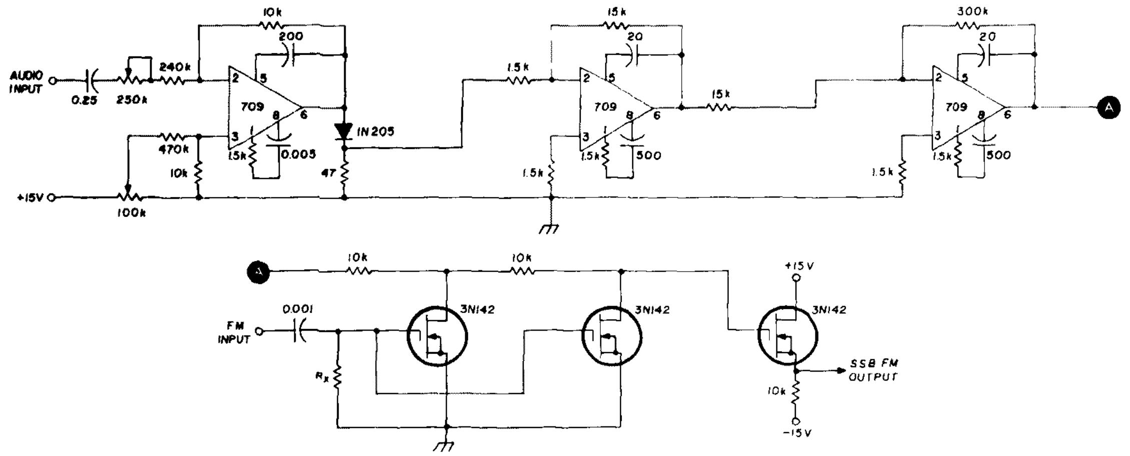


**fig. 4. Snider and Schilling's experimental "exponentiator" and amplitude modulator. A dual ±15 volt supply is used. Resistor Rx was not specified by reference 6. Note that this is an example only and is not intended for home construction.**

For those who are experimentally inclined, Snider and Schilling's article is recommended since it also contains schematic diagrams of a useful ssb fm amplifier, designed to produce 455-kHz ssb fm. Glorioso and Brazeal[4] give schematics for a discrete exponentiator using four transistors and also for a frequency multiplier. In fact, these two articles contain 90 percent of the constructional material which has been published in the open literature.

## bandwidth of ssb fm

Now that I have touched very lightly on the electronic details of ssb fm generation, I will turn to a question which was never answered to everyone's satisfaction; namely, what is the bandwidth of a typical ssb fm signal?

The answer is not necessarily "one-half the bandwidth of an ordinary fm signal," and in fact depends upon the modulating waveform.[7] This is because fm produces multiple sidebands for each modulating frequency, so the bandwidth can't be as easily defined as it is for a-m. Usually the bandwidth of an fm signal is considered to equal the spectrum space which contains some fixed percentage (often 90) of the radiated power. When sideband enhancement takes place in an ssb fm generator, sidebands which didn't contain enough power to contribute to the bandwidth of the parent signal may become important. If they do, the resulting ssb fm signal will be wider than one-half of the original signal. Just how much wider, or if the signal is always wider, has never been settled because as shown in fig. 5, the redistribution of relative sideband power is quite uneven.

A practical rule-of-thumb would seem to be that, first, single-sideband signals derived from wideband fm (modulation index greater than 0.4) are about two-thirds as wide as ordinary fm and, secondly, single-sideband signals derived from narrow-band fm are one-half as wide as nbfm. However, as the later discussion will show, there are many who disagree with this evaluation.

## effect of limiting

Recall that the act of amplitude-modulating the parent signal cancelled an unwanted group of sidebands; hence, the amplitude variations present in the ssb fm signal can be thought of as information about how the generator accomplished its goal. If that information is removed, it is equivalent to negating the influence of the amplitude-modulator at the generator so the original sidebands are all restored, and ordinary fm results. It follows that the act of subjecting ssb fm to hard limiting

restores it to ordinary fm, and it was this potential compatibility with standard fm broadcast receivers that first attracted the attention of the developers of the ssb fm technique.

As was mentioned earlier, if ssb fm were viewed as an end in itself, you would *not* use an ordinary fm receiver to detect it, but would go in the direction indicated by Kahn and Thomas[5] – detecting both amplitude and frequency changes for later processing. Nevertheless, virtually everyone who worked with ssb fm considered only ordinary (consumer) fm receivers because they saw ssb fm as a possible way to squeeze more stations onto the commercial fm broadcast frequencies. This is an important point to keep in mind, because when you take

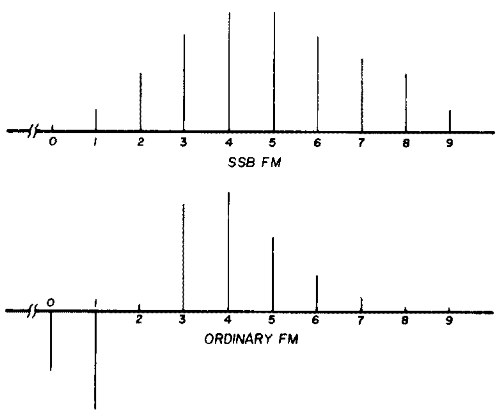


**fig. 5. Relative upper sideband amplitudes for ssb fm (top) and ordinary fm (bottom) versus sideband number, where zero indicates the carrier component. Both plots are for a modulation index of 5, but are drawn to different vertical scales. Top: Carrier component equal to unmodulated signal amplitude. Bottom: Carrier component 18% of unmodulated signal amplitude.**

only that viewpoint, you can easily "prove" that ssb fm is inferior to ordinary fm. The next section summarizes those arguments.

## the case against ssb fm

The bandwidth (however you choose to define it) of an ssb fm signal can depend so strongly upon the modulating waveform that very great differences in its bandwidth were obtained by several workers. Mazo and Saltz,[7] who assumed so-called Gaussian modulation, concluded that ssb fm was often wider than fm whenever the modulation index exceeded three. Hence, they declared that it was of little interest for such purposes as standard broadcasting. Kahn and Thomas[5] reached the opposite conclusion for the case of sinusiodal modulation, declaring that wideband ssb fm is half as wide as ordinary fm but that narrow-band ssb fm is 1.4 times as wide! Both of these papers were entirely theoretical, and neither of them contained supporting laboratory experiments.

On the other hand, Glorioso and Brazeal built a working system and agreed with Bedrosian's earlier conclusion that a narrower bandwidth is to be expected for ssb fm. In fact, the bandwidth rules given earlier in this article are based upon these author's findings, as well as those of Dubois and Aagaard; it is believed that they represent the current best choice among decidedly divided opinions. Nevertheless, it could still turn out that ssb fm is unattractive due to bandwidth limitations.

A more certain criticism of ssb fm arises when you consider the signal-to-noise ratio at a conventional discriminator: The amplitude variations inherent in the ssb fm signal make it more difficult to obtain full limiting, so that the threshold characteristics are poorer and the threshold signal-to-noise ratio is degraded. These problems become particularly troublesome at large modulation indexes, as does an accompanying problem of maintaining phase linearity in the receiver i-f strip.[6] For all of these reasons, Snider and Schilling expressed the opinion that ". . . ssb fm will not find broad practical application, since it is both difficult to generate and does not perform as well as fm does."[6]

## the case in favor of ssb fm

It must be repeated that the usefulness of any communications method depends upon the way in which it is configured. Certainly, if you want to broadcast compatible, high-fidelity ssb fm into the home of an audiophile, you have chosen an impractical goal. On the other hand, if you are interested only in communicating information to a *matched* receiver, ssb fm may indeed be attractive. For example, Glorioso and Brazeal held that, even in the wideband case, "Envelope detection, additional nonlinear processing, and filtering . . . offers threshold characteristics similar to conventional fm. . ."[4] Furthermore, they maintained that there is an accompanying saving in spectrum space.

In the narrowband case, ssb fm is readily detected with a conventional receiver and offers an improved signal-to-noise ratio over conventional fm once the detection threshold is exceeded.[4] Even though the threshold signal-to-noise ratio is still slightly poorer than that of conventional nbfm, it is believed that the loss is a very slight one or two dB.[4]

The accusation that ssb fm is "difficult to generate" depends, of course, upon your own viewpoint. Since Glorioso and Brazeal have shown that ssb fm can be generated at low frequencies and then multiplied to a higher frequency (and a higher modulation index), it is slightly simpler than ssb a-m, which must always be moved by heterodyning (mixing) techniques. Ssb fm is very much like ssb a-m in its requirements for generation and linear amplification, but these are routine requirements and constitute no serious difficulty. In short, the arguments in favor of ssb fm roughly counterbalance the negative opinions, and there does not seem to be sufficient information as of yet to make a go/no-go decision.

Particularly intriguing is the question of suppressed-carrier operation – a possibility which has not been discussed in the literature. Unlike conventional fm, it turns out that ssb fm contains a constant-amplitude component at the carrier frequency, so that carrier re-

insertion at the receiver appears to be feasible. This is not immediately obvious from the information given in the references, because each author chose to normalize his calculated sideband amplitudes in such a way that the carrier amplitude appears to decrease as the modulation index is increased. Furthermore, Glorioso and Brazeal show a plot of "carrier level" versus modulation index which, at first inspection, seems to imply that the carrier amplitude does change. In reality, the plot is for the signal level, which is something quite different. Anyone who decides to consult the references should keep these facts in mind.

The main point to be made here is that carrier suppression, though apparently feasible, has not been attempted. When you consider where ssb a-m might be if no one ever thought to suppress the carrier, it is natural to wonder if this is not the next logical step to take in testing the usefulness of ssb fm.

It should be mentioned that ssb phase modulation (ssb pm) may be more attractive to amateurs than ssb fm. The reasons for this are twofold:

**1.** In the case of ssb fm, the peak envelope power (PEP) increases with decreasing modulating frequency, whereas with ssb pm the PEP is independent of modulating frequency (see expression four in the appendix). Hence, ssb pm is not as likely to overdrive any following linear amplifiers at low audio frequencies.

**2.** An ssb pm generator can be built from fig. 3 by eliminating the audio integrator and replacing the vco with a true phase modulator. Therefore the circuitry is simpler.

Unfortunately, an ssb pm generator has never been explicitly discussed in the open literature so a potential designer is entirely on his own. However, this is not a very serious drawback because any existing ssb fm circuitry can be easily modified as indicated above.

## closing comments

During the preparation of this article, several amateurs were asked to review it and comment. It was an almost unanimous opinion that a complete schematic diagram of a working ssb fm transmitter, along with alignment instructions, ought to be included, and I assume that a number of readers will agree. However, it is my feeling that, in addition to being beyond the scope of a preliminary article on the subject, the actual hardware is probably illegal to use in the United States due to its use of *simultaneous* fm and a-m. If the FCC makes its proposed rule changes, then ssb fm may turn out to be legal without special permission. In that case, it would be useful to explore specific equipment after it has been given on-the-air tests, but so far, no one in the amateur community can do this and be sure that it is a lawful activity. Clearly, simultaneous fm and a-m has heretofore been thought of as a sign of sloppy hardware design, rather than a valid means of signal generation, and clarification is needed. By default, ssb fm is emission type F9 (there is no F3a or F3j classification for fm), and F9 emissions are not allowed on any amateur frequency.

It is hoped that readers of this article who have access to a good technical library will be sufficiently attracted to ssb fm to pursue it to the point of designing appropriate equipment. Certainly there is more to be done in the area, and amateurs can make some genuine contributions to the art.

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J.W. Couch, "Signal-to-Noise Ratio Out of an Ideal FM Detector for SSB FM Plus Gaussian Noise at the Input," *IEEE Transactions on Communications Technology*, 1965, page 591.

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## appendix

For those readers who wish more precise details, here are some mathematical relationships:

1. Single-tone ssb fm signal voltage $v(t)$: $v(t) = V_o \exp(\beta \cos \omega_m t) \cos(\omega_o t + \beta \sin \omega_m t)$

where:

$\beta$ = modulation index
$\omega_m$ = modulating angular frequency
$\omega_o$ = carrier angular frequency, and
$V_o$ = unmodulated carrier amplitude

2. Single-tone ssb fm sideband amplitude $A_m$

$$A_m = V_o \beta^m / m! \qquad m = 0, 1, 2, \ldots$$

where:

$m = o$ yields the carrier amplitude $A_o$

3. Peak-to-valley ssb fm envelope voltage, $V_{PV}$

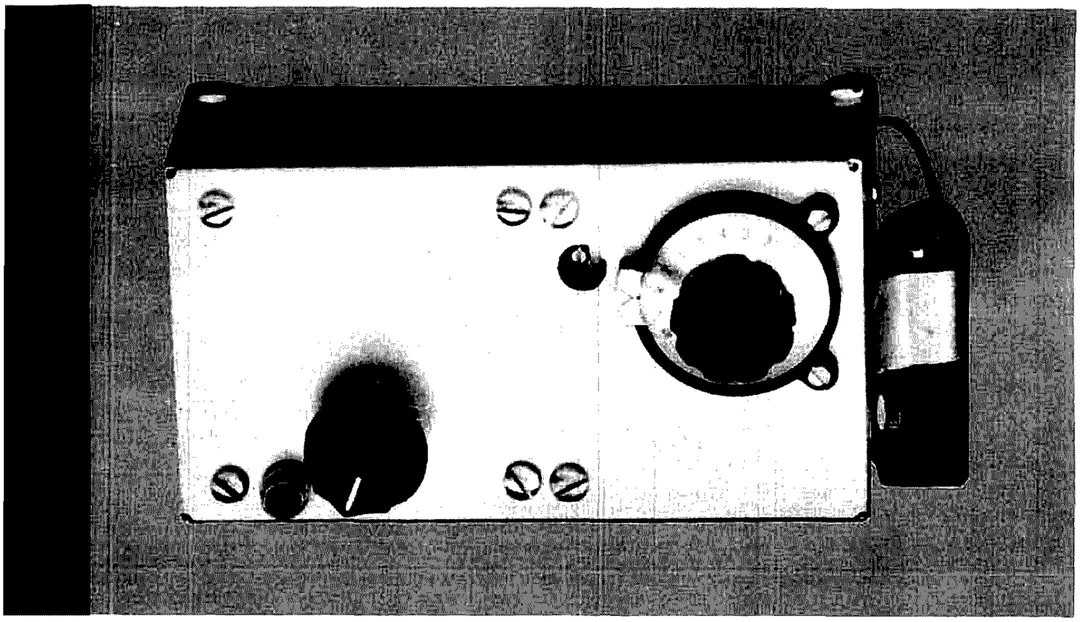
$$V_{PV} = 2V_o \sinh \beta$$

4. Peak-to-peak ssb fm envelope voltage, $V_{PP}$

$$V_{PP} = 2V_o \exp \beta$$

In all of the above relationships, the ssb pm case may be obtained by replacing $\beta$ with $\phi$, the maximum phase excursion in radians.

**ham radio**



# direct-conversion receiver

## for 40 meters

### Complete construction and alignment details for a mini receiver that can be built by beginner or old timer

Here's a tiny solid-state receiver with some great characteristics for headphone reception on 40 meters. The receiver has low-noise properties and anti-overload characteristics as good as those found in some of the better vacuum-tube receivers. Idling-current drain is less than 15 mA using a 9-volt transistor battery. For identification, I call it the DCM-1. It uses an rf jfet, a vfo jfet, an audio jfet, and an op amp final audio stage. Reception is best on CW, but single sideband can be received satisfactorily when the band isn't too crowded. A miniature resonant whip antenna can be used for satisfactory reception from all over the United States.

This project was tailored with the beginner in mind. The receiver is ideal for monitoring 40-meter novice stations without requiring an outdoor antenna. The parts list is quite complete and includes many substitutions that will perform equally as well as those given as a first choice. I've included technical expanations for completeness, but don't be alarmed if at first you don't understand this material when you attempt to build this receiver. After all, we all have to start somewhere, and most of us aren't going to make the big league anyway. I was once a beginner and still don't have all the answers.

If you follow the construction directions you'll have a piece of equipment any amateur would be proud to own. A block diagram of the receiver is shown in fig. 1.

### product detector and rf circuit

Many references have been made to direct-conversion receivers in the amateur literature. The heart of this receiver is a combination of three jfets in a Y connection such as used in a differential amplifier. This combination is almost the fet equivalent of the well-known RCA bipolar IC, type CA3028A, which has been used in product-detector circuits[1] among its many other applications. One advantage of the fets is that a much lighter load is presented to the vfo than when using bipolar transistors, so no buffer-amplifier is needed. Another advantage is that a better noise figure seems to prevail. Still a third advantage is low current drain. In fig. 2 it's obvious that the total current drain is only that drawn by the third transistor in the Y-connection, Q3. In the case of the prototype receiver, this current was only 8 mA.

In fig. 2, Q1 and Q2 are the matched jfets and Q3 is the series-connected rf amplifier. The variable frequency oscillator (vfo), which functions as the local oscillator (lo), is fed to gates G1 and G2 of Q1, Q2 in push-pull

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through conventional trifilar-wound broadband ferrite transformer, T1, with a 4:1 impedance ratio.[2] Q1 and Q2 switch the rf drain current of Q3, producing a mixer output at a frequency difference between that of the vfo and the received signal, which is in the audio range. The audio output is obtained through push-pull transformer T2, whose secondary is resonated to approximately 750 Hz by capacitor C2. I found it essential to use zener diodes CR1 and CR2 across the Q1, Q2 output, otherwise high-voltage switching transients generated in T2 can burn out the twin fets.

The rf signal is fed from the antenna jack to Q3 *through a double-tuned resonant circuit* consisting of L1, L2. Each inductor of L1, L2 has an unloaded Q of about 220. The two resonant circuits are coupled through C7, a very small capacitor, whose value was found by experiment. Capacitor C7 is merely a twisted pair of plastic-covered solid no. 22 AWG (0.6mm) wires. Resistor R2 was set at 3300 ohms in the prototype, but should be adjusted as described below.

The original circuit was designed for use with the Siliconix E421 dual fet in the Q1, Q2 position and the E304 in the Q3 position. Both are marketed under a different number by Calectro (at least as of this writing). Using the dual fet, the circuit should be sufficiently well balanced in some applications not to require equalization of Q1, Q2 gains. For cases where equalization is desired, or where individual fets are used, the alternative connection at points X, Y, and Z in fig. 2 should be followed, requiring a slightly different circuit-board layout, which is discussed later.

## vfo (local-oscillator) circuit

The circuit of fig. 3 is a straight-forward series-tuned Colpitts circuit,[3] also known as the Clapp circuit. Biasing diode CR1 was chosen as a germanium diode,

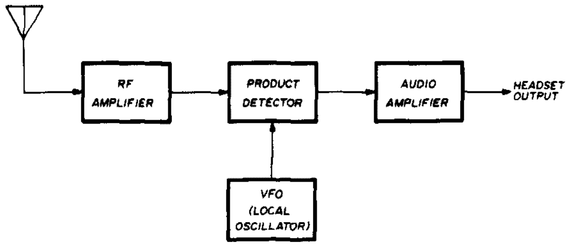


**fig. 1. Block diagram of the direct-conversion 40-meter mini receiver.**

because in past experiments I found that the oscillator had lower current drain than when using a silicon diode, but the choice is probably not critical. In the Clapp circuit, the impedance is low across feedback capacitors C5 and C6. Therefore it was possible to mount the tuning elements on a separate board from the transistor circuitry, locating the former conveniently close to the variable capacitor and connecting them to the latter through a 50-ohm miniature cable.

Rf chokes L3 and L4 are not critical. In my case they were wound on small ferrite cores as described in **table 1**. Small chokes, such as Millen subminiatures of 10-15 $\mu$H would be satisfactory. L2 is a Radio Shack 2.5 mH choke, but a small choke of the order of 50 $\mu$H would be satisfactory. Resistor R5 was 3900 ohms in the prototype, but might best be adjusted for the lo output that affords the highest detector audio output, as described later.

The vfo tuning inductor, L1, is a high-Q toroid. The tuning capacitor, C3, is a miniature 365 pF air-dielectric broadcast unit. Since this capacitor tunes through a large range, it was necessary to swamp its effect with fixed capacitor C4, which was 470 pF in the prototype. The

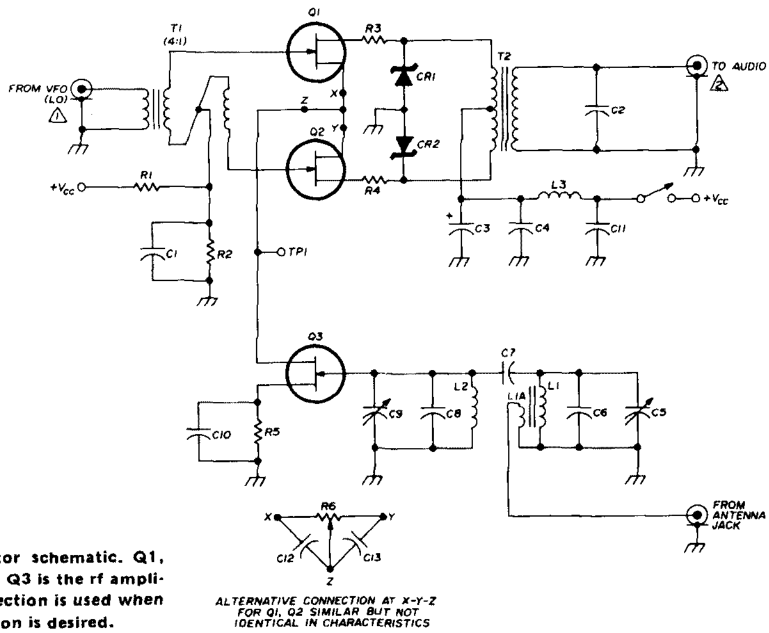


**fig. 2. Product-detector schematic. Q1, Q2 are matched jfets; Q3 is the rf amplifier. Alternative connection is used when Q1, Q2 gain equalization is desired.**

**table 1. Parts list for the DCM-1 40-meter receiver. All resistors are ½ W, 10% tolerance. Alternative parts choices are also given.**

**product detector:**

| | |
|---|---|
| C1, C4; C10 through C13 | 0.02 μF, 25 V disc ceramic |
| C2 | 0.05 μF, 100 V mylar |
| C3 | 100 μF, 15 V miniature electrolytic, radial leads |
| C5, C9 | 5-30 pF ceramic trimmer |
| C6, C8 | 100 pF polystyrene or silvered mica |
| C7 | coupling gimmick, twisted pair of solid no. 22 AWG (0.6mm) plastic-coated wire, 3½ in. (89mm) long |
| L1 | 30½ turns no. 24 AWG (0.5mm) on T50-6 toroid core |
| L1A | 2½ turns no. 24 AWG (0.5mm) wound over ground end of L1 |
| L2 | 31½ turns no. 24 AWG (0.5mm) on T50-6 toroid core |
| L3 | rf choke, 2.5 mH (Radio Shack 270-1713) smaller unit OK |
| Q1, Q2 | twin rf fet in one package (Siliconix E421 or Calectro Equivalent). Alternatively, any two hf-vhf n-channel jfets reaonably similar and used with alternate source-current balancing circuit. |
| Q3 | Siliconix E304 or Calectro equivalent. Alternatively, a third jfet similar to Q1, Q2 may be substituted. |
| R1 | 47k |
| R2 | Set for dc voltage source-to-drain across Q3. Use high-impedance voltmeter. Prototype receiver value was 33k |
| R3, R4 | 10-22 ohms; not critical |
| R5 | 47 ohms |
| R6 | source-current balance pot, 100 ohms (miniature Bourns type) |
| T1 | 13 turns no. 30 AWG (0.25mm) trifilar wound on Indiana General ferrite core CF102-Q1 or Amidon T25-6 core. Impedance ratio, 4:1 |
| T2 | Push-pull audio transformer, Calectro D1-722 or equivalent |
| CR1, CR2 | 18V, ½ W zener diodes |

**vfo (local oscillator):**

| | |
|---|---|
| C1 | small 180 pF polystyrene or silver mica (see text) |
| C2 | 5-30 pF ceramic trimmer |
| C3 | 365 pF broadcast-band capacitor |
| C4 | 470 pF polystyrene or silver mica (see text) |
| C5, C6 | 560 pF polystyrene or silver mica |
| C7 through C11 | 0.02 μF, 25 V ceramic |
| CR1 | any germanium receiving diode |
| L1 | 49 turns no. 24 AWG (0.5mm) on Amidon T50-6 toroid core |
| L2 through L4 | choke, 18½ turns no. 26 AWG (0.3mm) on CF102-Q1 ferrite core. Not critical. Alternative: fill T25-6 toroid with no. 32 AWG (0.2mm) wire |
| Q1 | any good hf-vhf n-channel jfet (Siliconix E304 or equvalent) |
| R1 | 47k |
| R2 | 330 ohms |
| R3 | 4.7k |
| R4 | 10 ohms |
| R5 | 3900 ohms in prototype; see text |

**audio amplifier:**

| | |
|---|---|
| C1, C2 | 0.01 μF, 50 V |
| C3, C5 | 25 or 30 μF, 15 V miniature electrolytic, radial leads |
| C4, C8 | same as above, but 50 μF |
| C7 | same as above, but 10 μF |
| C6 | 0.1 μF mylar |
| C9 | 0.005 μF mylar |
| Q1 | n-channel jfet (2N3819, HEP801 or similar) |
| R1 | 100k miniature volume control with on-off switch |
| R2 | 43 ohms |
| R3 | 27k |
| R4 | 22ks, |
| R5 | 2.7k |
| R6 | 2.7 megohm |
| R7, R8 | 5.6k |
| R9 | 220 ohms |
| U1 | type 741 op amp |

**miscellaneous:**

aluminum box, 5¼ x 3 x 2-1/8 (13x8x5.4cm). Bud CU3006A or Radio Shack 270-238

phone jack (size depends on plug size)

phono jack for antenna connection

dial drive, Calectro E2-744, diameter 1½ in. (38mm)

battery and battery terminal. Use heavy-duty 9-volt unit

4 spacers, 3/4 in. (19mm) tapped for 6-32 (M3/5) to support product detector panel

2 spacers, ½ in. (12.5mm) tapped for 6-32 (M3/5) to support vfo tuner panel

4 spacers, ½ in. (12.5mm) to accomodate 6-32 (M3/5) screws to support rear panels

2 or 4 4-40 (M3) binder-head screws, ¼ in. (6.5mm) long to mount variable capacitor

8 6-32 (M3/5) binder-head screws, ¼ or 3/8 in. (6.5mm or 9.5mm) long

4 6-32 round-head screws to mount rear panels

2 no. 4 sheet-metal screws 3/8 in. (9.5mm) long to secure dial drive to front panel

lockwashers: 12 6-32 (M3/5) and 4 4-40 (M3)

1 circuit board, fiberglass-epoxy, 2½ x 2½ in. (64 x 64mm), copper foil one side

3 circuit boards as above but 2½ x 1-1/8 in. (64 x 29mm)

3 ft (1m) RF-174/U coax cable

10 ft (3m) hookup wire no. 22 AWG (0.6mm) solid

2 transistor sockets

1 14-pin dip IC socket

overall capacitance of the combination was too high, so C1 was inserted in series (180 pF in the prototype). Capacitor C2, a 5-30 pF ceramic trimmer, sets the tuning at a reference frequency and dial position using a wand applied through a hole in the front panel. The prototype oscillator current drain was about 3 mA.

## audio circuit

This circuit, fig. 4, is simple yet is the fourth one tried. The small broadcast transistor radios of the pre-IC days employed transformer-couple push-pull Class B output stages to obtain reasonably undistorted speaker out-

put at only 200 mW. However, I noted that the overall drain of such a receiver was 7-8 mA, increasing to 16-17 mA on audio-modulation peaks. I found that even at comfortable headphone volume, a single output transistor produced intolerable distortion. An audio fet, Q1, is biased to draw only a few tenths of a milliampere. Its output feeds a type 741 operational amplifier, U1, which works satisfactorily even on a 9-volt battery, drawing about 2.5 mA. The overall receiver current increases at most about 2 mA on loud signals with the volume turned up. The circuit is the same as that of reference 4, except that resistor R6 was increased to 2.7 megohms from 0.82 megohms, affording about 10 dB more signal gain.

## construction

**Circuit boards.** **Fig. 5** is a layout of the product-detector board for the twin-fet circuit without balancing adjustment. It shows component location and drilling points. **Fig. 5** is an etching template. **Figs. 6A** and **6B** are for the product detector where a balance control is desired, whether a twin fet or individual fets are used. Mounting holes for adjustable ceramic trimmers C5 and C9 may have to be changed, depending upon the style available.

**Fig. 7** shows etching templates for the small panels: **fig. 7A** the vfo oscillator; **fig. 7B** the vfo tuner; and **fig. 7C**, the audio amplifier. **Fig. 8** shows component and drilling locations for these panels. Note that the audio-amplifier layout is for the 14-pin DIP package socket. For the TO-99 package, stretch the leads to the proper hole numbers, as indicated in fig. 7, and solder. A separate layout is needed for the mini-DIP package. Use care to identify leads from the top of the IC, if this is what the manufacturer's diagram specifies. All boards should be of fiberglass-epoxy resin, such as G-10, with fiberglass paper as a possible substitute, and should have copper foil on one side.

In the type of construction I favor, the conductors are placed like islands surrounded by insulating material, all in a grounded matrix of copper. If this language

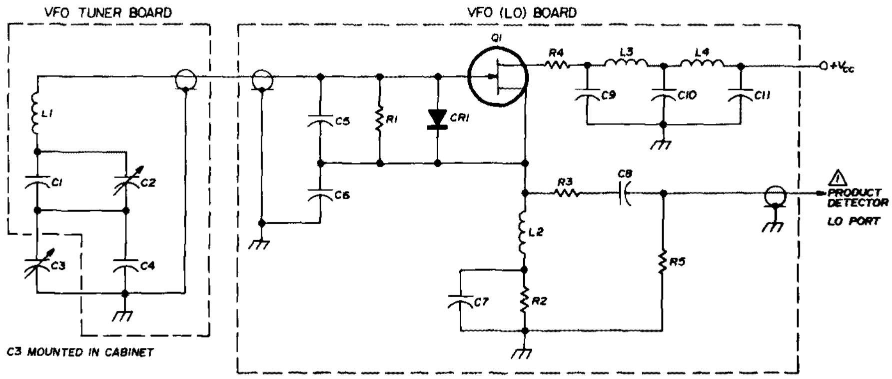


**fig. 3. Vfo (local oscillator) schematic. A simple series-tuned Colpitts circuit is used for low current drain.**

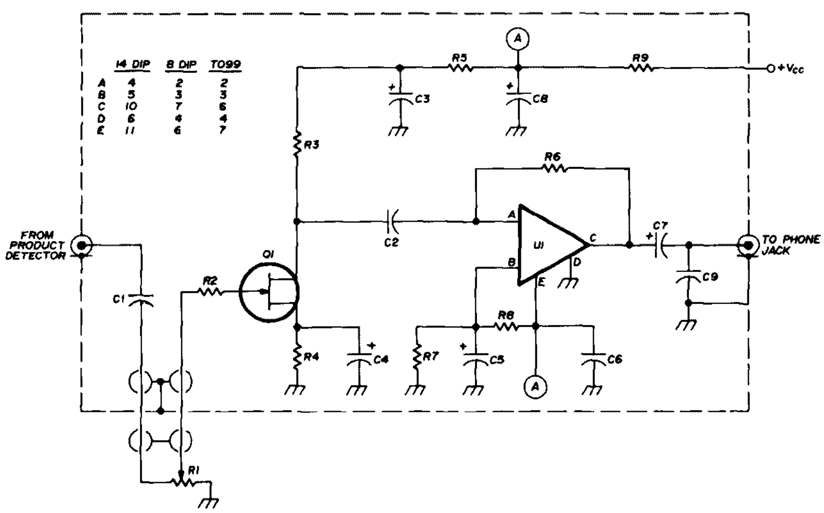


**fig. 4. Audio-amplifier schematic. Circuit is similar to that in reference 4. R1 is mounted on front panel.**

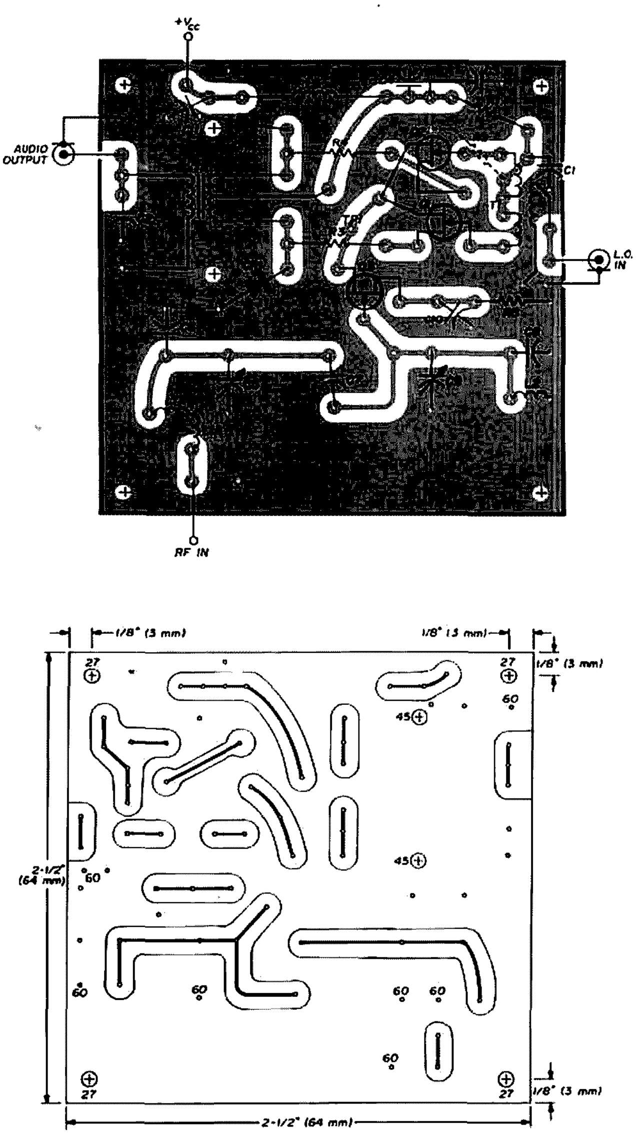


**fig. 5. Product-detector layout using twin fets, above, and its etching template, below.**

seems complicated, you'll understand by looking at the etching figures. The method takes a little more artwork but actually consumes much less etching fluid than if conductors were laid down individually. Moreover, a surrounding ground plane affords optimum shielding.

**Drilling and soldering.** Drills should be of the hardest possible steel, since fiberglass is an abrasive material. At least one source of such drills in quantity at reasonable prices has been advertised.* All small holes not specified in the diagrams are to be drilled with a no. 65 (0.9mm). A stable floor- or table-mounted drill press will work satisfactorily. Some of the holes marked for no. 60 (1mm) are obviously (from inspection of the board diagrams) to be used for grounding the braid on cables terminating adjacently, so that they may need to be still larger, perhaps no. 54 (1.4mm). Before mounting components, check all islands for shorts to ground and remove any with a knife or scratch awl.

Use care not to apply too much heat in soldering; conductors etched on the best boards can strip off! Boards made of fiberglass and paper, and others with more of a paper composition, tend to smoulder with excess heat. A pencil-type iron for circuit-board work is desirable. Use a quick, firm push with a clean iron tip into the joint, bringing the solder up simultaneously and withdrawing quickly. For best results, the copper should be cleaned immediately before soldering to inhibit corrosion. Plating the finished board with tin or silver will improve soldering effectivity. (A chemical tin-plating kit is sold by Burstein Applebee). The small battery-operated soldering irons are great for getting into small places, and there is an advantage in not dragging a cord around.

After a board has been completed care must be taken to clean off any rosin that bridges the insulating space between conductors and ground. This is important on high-impedance circuits and if the equipment is to be used in a humid climate.

A question arises as to whether to solder transistors and ICs in place or to use sockets. Space and cost are saved by not using sockets, but plugging semiconductors in and out makes it a lot easier to use cut-and-try procedures. In the product-detector board, I soldered the units in place, mainly because the E421 fet would have required an awkward socket wiring arrangement; the other boards are laid out so that sockets may be used if desired. Protect semiconductor leads from heat when soldering and desoldering.

Circuit changes and repairs are not too difficult if you remember not to apply too much heat and to heatsink semiconductors whenever necessary. If a removed component leaves a hole plugged with solder, the solder blob may be removed easily with rosin-coated copper braid sold under the name of *Solder Wick.*

## testing and adjustment

**Product detector and rf amplifier.** The first test and adjustment procedure applied to the product detector is to determine the equality of dc voltages across the upper transistor pair and the lower transistor. Voltages should be measured using test point TP1 (fig. 2), which is a short tinned wire to which small clips may be attached. Measurements should be made with a high-impedance voltmeter. No vfo voltage should be applied to the lo port of fig. 2. Leads to a 10k pot can be temporarily attached in lieu of R2. The correct setting of the pot determines the value of R2 to the nearest fixed value available. Equalizing the dc voltages is desirable to avoid running Q1, Q2 at too-low voltage. Under these conditions, an fet will be driven into regions of the operating characteristics considerably below pinch-off, resulting in excessive distortion. This was especially true with the E421s and E304s tested, where the zero-bias saturation drain current, $I_{dss}$, was not reached until the drain voltage was close to 6 volts; i.e., $V_p$, the pinch-off

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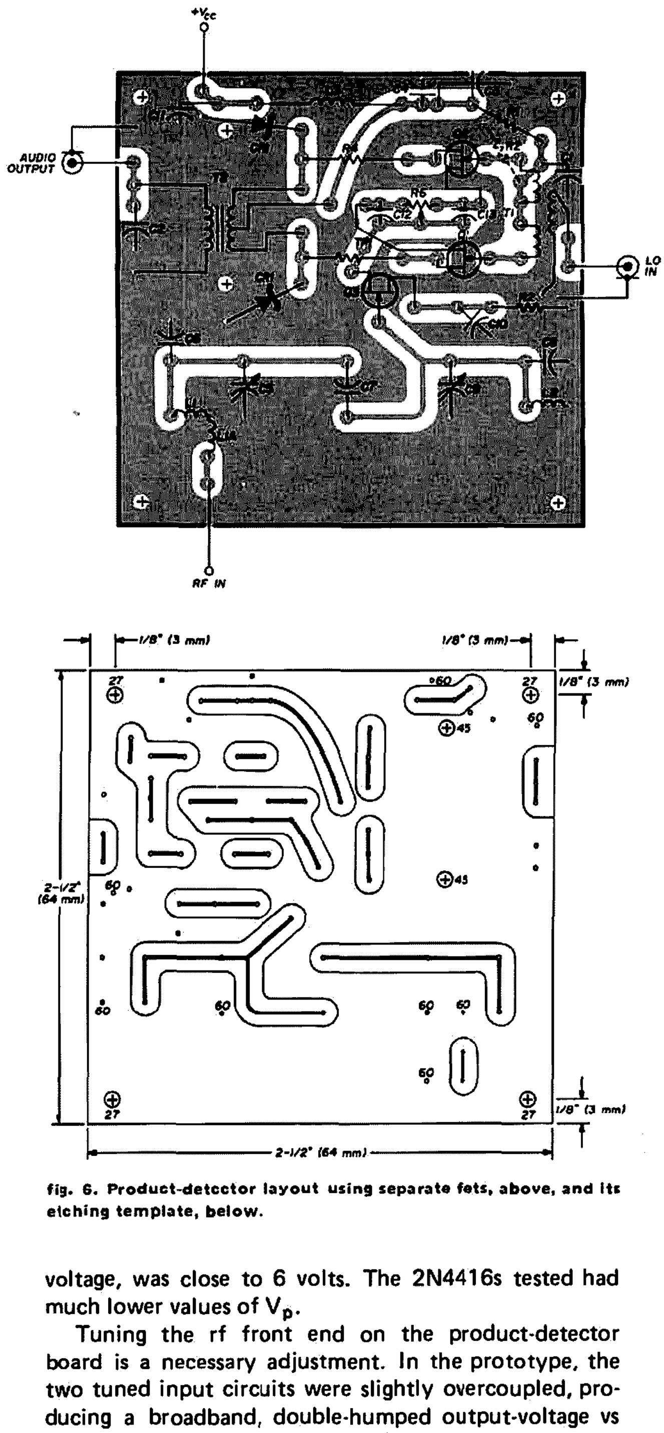



fig. 6. Product-detector layout using separate fets, above, and its etching template, below.

voltage, was close to 6 volts. The 2N4416s tested had much lower values of $V_p$.

Tuning the rf front end on the product-detector board is a necessary adjustment. In the prototype, the two tuned input circuits were slightly overcoupled, producing a broadband, double-humped output-voltage vs frequency response, covering 7.0-7.3 MHz. The mixer was operated again without vfo input. In my tests the output of Q3 was measured at TP1 with an rf probe attached to a vacuum-tube voltmeter. Q3 output can also be measured with an oscilloscope probe it the oscilloscope vertical amplifiers will pass 7 MHz. Yet another way is to fashion a probe with a piece of coaxial cable. A small capacitor is placed in series with the center conductor and touched to TP1. At the probing end of the cable the shield should be grounded to the board with a short clip lead. The other end of the cable should go to a receiver, and the output is read on an S-meter.

Still another way to read Q3's relative output is to operate the mixer as a mixer and read the audio output at the beat-note frequency that gives maximum audio output. Many test voms have an ac response up to 1000 Hz. One could be used at the 741 output with the headset disconnected. This method requires a minimum of test equipment but is a little more critical in adjustment.

With the relative-output technique determined, the actual adjustment of the tuned circuits is best accomplished by the following method. Use a signal generator or other equivalent shielded source with a means of attenuation. Attach small clips to short leads of a capacitor of about 470 pF. Tune the signal generator to the center of the desired frequency range and feed its output into the antenna input. Short-circuit L2 and its capacitors with the test capacitor. Force a large signal through and adjust C5 for maximum Q3 output, making sure Q3 is not overloaded. Next, clip the capacitor across L1 and tune C9 for maximum. Always make sure that the maximum occurs in two places as the capacitor is rotated through 360 degrees, otherwise the circuit must be altered to produce resonance.

After both sides are resonated, the auxiliary shunting capacitor is discarded. Then it should be possible to measure output with a small-signal input, and it will be necessary to reduce the input to prevent overload. As frequency is changed the output should be a double-humped curve if C7 is large enough to cause over-coupling. If C7 is reduced to a value equal to or less than that needed for critical coupling, the curve will be single humped at the original frequency of tune up. A too-small value of C7 will make the front end resonance curve too narrow for some purposes, and as C7 is made still smaller, the signal-to-noise ratio will deteriorate.

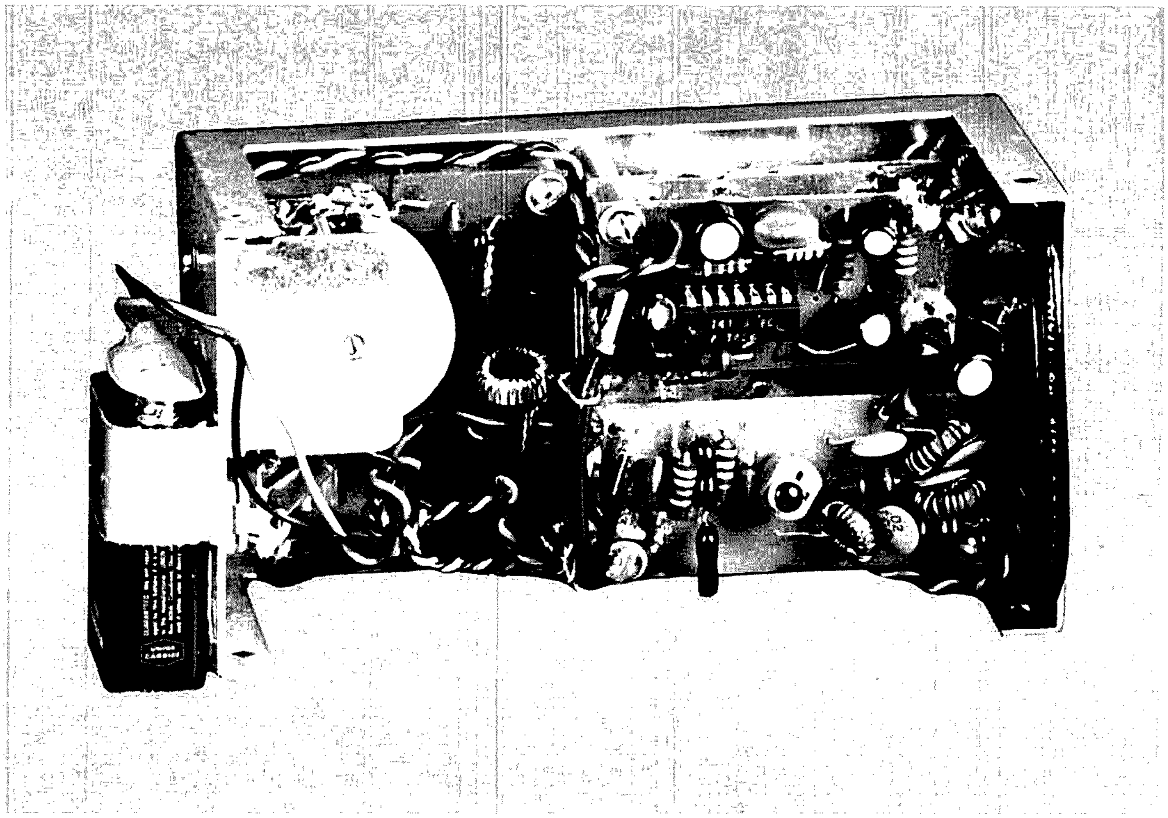
At this point it's evident we're going to have to test several units in conjunction with one another. It should be observed from the diagrams that all rf and audio leads connecting the boards to each other and to the cabinet are coaxial cables, whereas dc leads are pairs of wires braided together. A ground wire braided with battery leads affords some shielding. In the final layout, the units using coaxial cables should be joined by miniature cable lengths just long enough to enable servicing the receiver with boards removed, and the same applies for the dc leads. Earlier, somewhat longer pieces of coax can be used for convenience in bench-testing the newly built units.[5] Temporary dc leads are no problem. It's a good idea to use a current-limited power supply with a milliammeter in series with the line to any new unit to be tested. Gradually bringing up the voltage lets you know whether or not the current drain is reasonable.

**Vfo (local oscillator).** Vfo alignment is fairly straightforward, although the variable capacitor presents some problems. Before aligning the vfo, it's necessary to have the variable capacitor and the vfo tuning board fastened

in position in the aluminum box, so this procedure is described with reference to **fig. 9.** (Drilling and removal of burrs should have been done before mounting was started).

The prototype receiver was designed to use a small, imported 1-1/2 in. (38mm) dial drive with an 8:1 gear ratio, turning a miniature 365-pF air-dielectric capacitor. Difficulties arise from the fact that sizes and mounting-hole positions may vary in different shipments of the same units. The worst problem, however, is in aligning the capacitor shaft with the dial-drive collar. Even if the shaft could be perfectly aligned, the hole in the collar would probably be a poor fit for the shaft. In the prototype I hack-sawed the capacitor shaft to the proper length and drilled and tapped the collar and the shaft at right angles for two 4-40 (M3) screws. The result was a capacitor that doesn't rotate quite all the way to the closed position (the shaft slipped while drilling). Worse, some drive eccentricity occurred. With the capacitor screwed tightly to the cabinet side, the lateral pressures on the shaft were enough to cause plates to short circuit. The solution was to loosen the capacitor mounting screws and ground the capacitor frame to the vfo tuner board with a short piece of braid. *Solder Wick* was used as the braid. Eccentricity was still present, causing some annoying backlash during frequency calibration. A better form of dial drive seemed possible but could not be developed in time. Note that in rotational alignment, before drilling and tapping for set screws, the dial should read approximately 0 with the plates fully meshed.

The tuning-capacitor plates are quite close, so *great care* should be used to prevent metal chips from lodging between the plates and in the ball bearings. While drilling or filing, vulnerable places should be covered with masking tape. When drilling into a cabinet or chassis that already contains boards and components, it's frequently beneficial to stuff areas to be drilled with small pieces of paper towling or rag. Chips can be removed from odd corners in the same way that small hardware can be removed from tight places: wrap a piece of masking tape with the sticky side out at the end of a pencil and probe.



**Interior view of the direct-conversion receiver. Audio board is at upper right, with vfo board mounted below. Printed-circuit layouts are shown in fig. 7.**

After mounting the capacitor and dial drive, the vfo tuner board should be mounted, foil side down, on two ½-inch (12.5mm) threaded spacers. A short piece of solid, tinned no. 18 or 20 AWG (1.0 or 0.8mm) wire should protrude from the top of the board, to be soldered to the stator section of C3. See **figs. 7B, 8B,** and **9.**

Now let's resume the vfo test. One sure indication that the vfo is oscillating is that, as the dc voltage is raised, the current will increase until oscillation occurs. When this happens, the meter will jump to a lower value

of current. When the vfo is oscillating and voltage is reduced, the current will jump up when the oscillator quits oscillating, but at a lower voltage than where oscillation started as voltage was raised. Never plan on operating an oscillator at a supply voltage below that required to start oscillation, or even close to that upper point on the high-voltage side.

When the oscillator is working and feeding the product detector, the lower limit of the desired tuning range should be set by turning the dial to, say, 5 or 10, allowing some leeway for oscillator drift. Then frequency-set capacitor C2 should be adjusted for this frequency as received on a receiver. If it's not possible to hit the desired frequency and bandspread, remember that increasing C4 and decreasing C1 together increases the bandspread, but increasing C4 lowers the frequency and decreasing C1 raises it, so you have considerable leeway in setting the frequency and bandspread. For instance, to cover a limited portion of the 40-meter band (like the novice band), you might first try reducing C1 to 120 or 150 pF and increasing C4 to 510 or 560 pF.

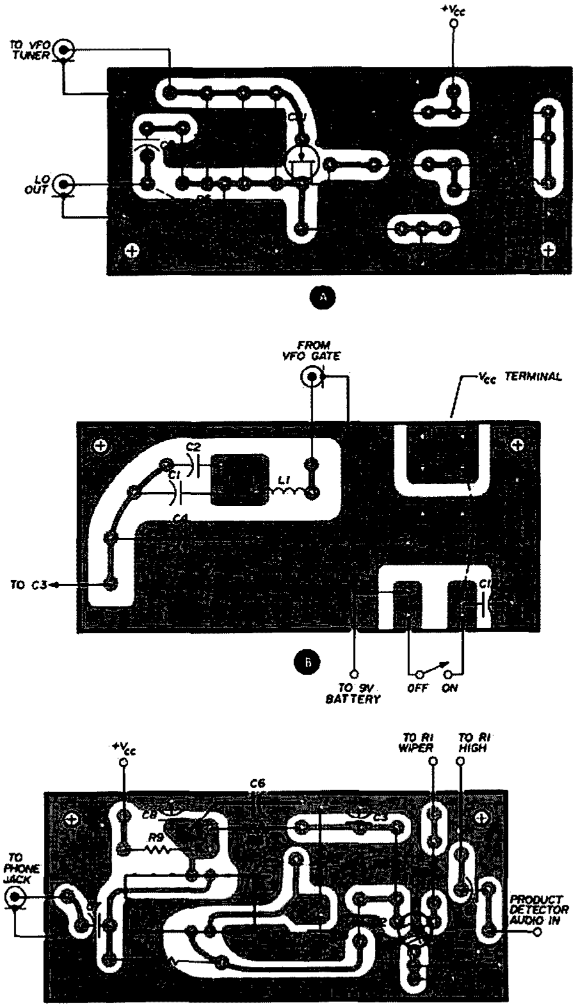



fig. 7. Component placement and etching templates for the small panels. Vfo oscillator board, vfo tuner board, and audio-amplifier board are shown in A, B, and C respectively.

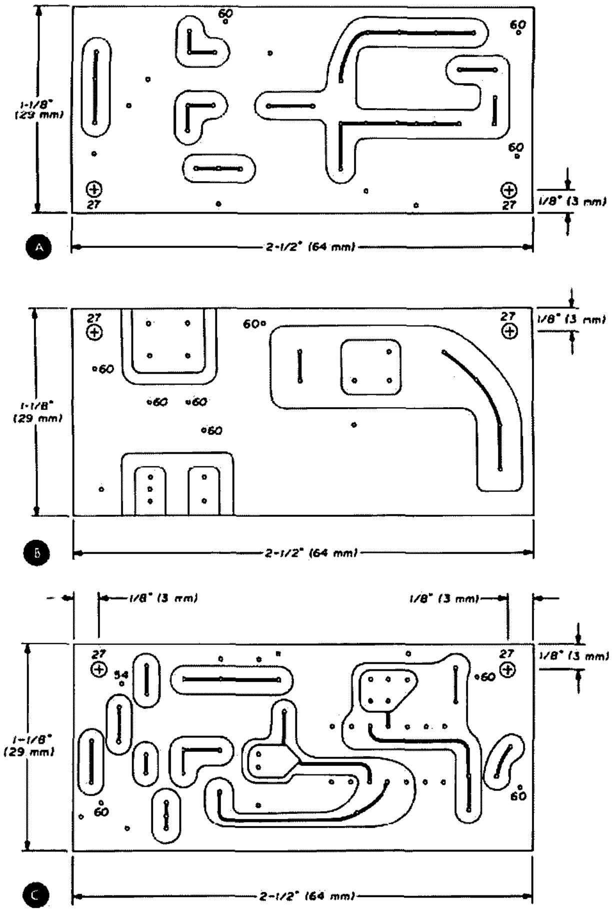



fig. 8. Circuit-board layout and drilling locations for the small panels. Vfo oscillator board is shown at A; vfo tuner and audio boards are shown in B and C.

The vfo local oscillator output voltage can be adjusted by varying R5. The local oscillator drive voltage to the product detector is not especially critical, but values of R5 in fig. 3 can be chosen by experiment for maximizing audio output. In my case 3900 ohms was found to be the best, producing an input voltage of 0.9 volts rms to each gate of Q1, Q2.

**Audio amplifier.** Audio board construction is straightforward, although components are a little crowded. Initial tests can be made with the volume control, R1, hanging loose. Although the output of the product-detector audio transformer is at high impedance, the 50-ohm shielded cables to the 100k volume control can be used, because they contribute in part to the capacitance that resonates the audio transformer secondary (i.e., at audio frequencies, a short length of cable is merely a small capacitance). Using these cables is necessary to prevent the high-gain audio circuit from self oscillating under certain conditions.

Final assembly into the small cabinet should be done by reference to fig. 9, and is much easier than it looks; it is a gradual process of fitting panels and shortening leads and cables. The 365-pF variable capacitor and the vfo tuner board have been previously mounted. After the volume control and antenna jack are mounted, the four 3/4-inch (19mm) threaded spacers are mounted to the rear end of the front panel. The product detector panel is then laid in position against these spacers. The foil side must face the rear, the bulky components occupying the space between the epoxy side and the cabinet front.

The audio and vfo boards are positioned behind the detector. The audio board mounts on top and the vfo board below, with the foil sides facing inward. They are spaced from the product detector by four 1/2-inch (12.5mm) spacers that are not threaded but pass 6-32 (M3/5) screws. The whole assembly is then secured by four 3/4-inch (19mm) 6-32 (M3/5) screws, which go into the original four threaded spacers, as shown in fig. 9. Lockwashers are used at front and rear.

The cables and dc wire pairs can be tucked into the space between the spacers and the box at the edges. Dress the wires and cables away from tuning inductors and the high-voltage side of all rf circuits.

The $V_{cc}$ wires should terminate at the $V_{cc}$ terminal, located for convenience on the vfo tuner board, fig. 8B. The ground wires can be soldered in the appropriate adjacent holes on the ground-foil area. A twisted pair should go to the on-off switch on the volume control. The 9-volt battery connects to a standard snap-on terminal with red and black wires, which go to points indicated in fig. 8B. Originally, the battery was forced between the variable-capacitor frame and the cabinet cover, but the pressure caused the capacitor to short in some positions. The battery was finally mounted on the outside of the cabinet with wires passed through a notch on the cover plate.

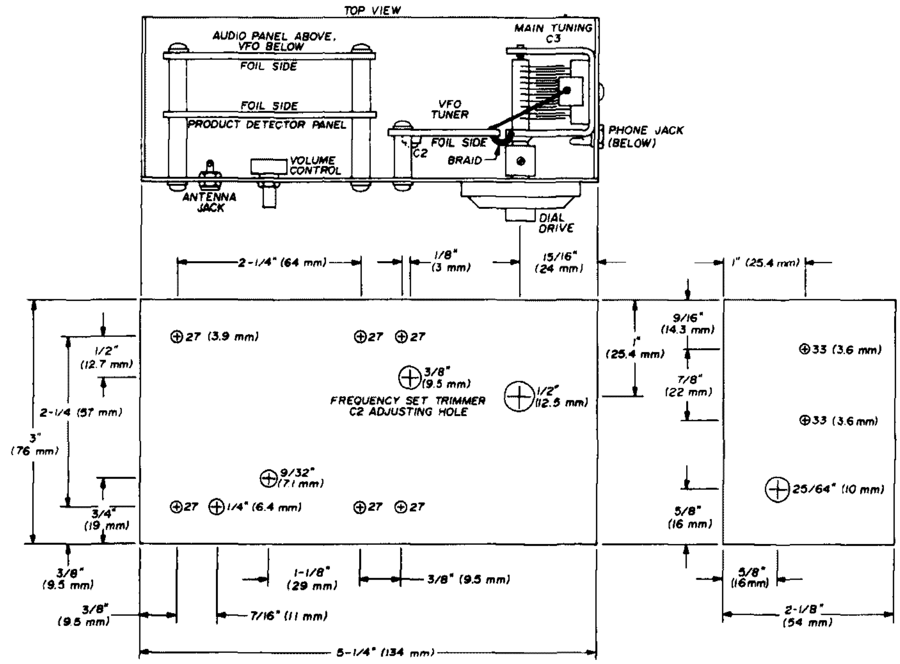



**fig. 9. Cabinet layout showing drilling and component-mounting details.**

## a miniature antenna

The receiver was designed to work with a 50-ohm outdoor antenna, but excellent results can be had with small antennas for indoor and outdoor use. Even a piece of wire strung up in the room and connected to the antenna jack will bring in many signals. A small, resonant loaded-whip antenna was designed that works very well for outdoor portable use and also indoors when not too near pipes, wires, heat ducts, and flourescent lights.

Fig. 10 illustrates the antenna. A section of RG-58/U coaxial cable with a phono plug at the lower end is strapped to a 3-foot (1m) piece of dowel. The dowel can be mounted in a board as a strand or fastened to the receiver with cable clamps, but in the latter position, there may be some chance of the lightweight reciever overturning. A fiberglass rod could also be used.

A small insulating board of fiberglass or phenolic is mounted flat to the top of the dowel. On one side is mounted a slug-tuned coil form such as J.W. Miller no. 43A001CB1, with the winding space filled with no. 28 AWG (0.3mm) enamelled wire. The end nearest the mounting nut is fastened to the cable center conductor. The shield at this point is not connected anywhere. The hot end of the coil is connected to a pair of galvanized-wire "antlers" spead horizontally from a screw on the other side of the board. Sections of hard-drawn tinned

busbar would be even better. The antlers provide top loading. At resonance, the antenna has a resistive impedance near 50 ohms, but it would be difficult to determine what part of this resistance is due to losses.

Approximate resonance can be found with a grid dipper and final resonance determined on background noise or with a signal generator on a small antenna. Tuning, of course, is with a tuning wand, keeping your body, phone cord, and other wiring as far away as possible from the system. A lighter weight antenna could be made using a toroid core on a T50-6 form, adjusting the turns to something near desired resonance. The slug-tuned coil can be protected by a plastic pill bottle.

## summary

One outstanding feature of this receiver is its large dynamic range. With a signal inserted directly at the gate of Q3 in the prototype, the output was linear with input up to a 1-dB saturation fall-off[6,7] at 850 mV rms input, representing an output audio voltage of 8.5 volts rms at about 750 Hz into the 100k volume-control load. Since the audio transformer has a turns ratio of about 2.24, this represented a push-pull drain voltage of 8.5/2.24 = 3.8. In the more linear portion of the characteristic, at lower inputs, the overall voltage gain was 13.3, or 22 dB (from the Q3 input to the transformer output). Of course, Q3 may have a gain of the order of 20 dB, *leaving* the *conversion* gain of switching pair Q1, Q2 small, but much better than if diodes had been used.

The double-tuned high-Q rf input has a voltage gain of 50, or 26 dB, so that the signal gain from the antenna jack to the audio-transformer output was 22+26, or 48 dB. Additionally the noise figure appeared to be very good. It was estimated that only about 0.2 mV input was needed for a 10-dB signal-to-noise plus noise ratio. Of course, you'd want to use the full resonant gain of the front end when using the small antenna. With a large antenna, you might be concerned that the saturation voltage is the original 850 divided by 50 (gain of the resonant circuits), or only 17 mV. Yet in practice, with an excellent outdoor antenna, I've failed to notice any great blocking of the receiver by strong signals – and here it must be remembered that the receiver has no agc circuit. When signals are too strong, merely turn down the audio volume. If blocking is a problem, an rf attenuator can be switched into the antenna input.[8]

The most obvious problem in this receiver is that the audio beat note appears on both sides of zero beat and selectivity is poor. DX can be received well when nearby stations are skipping over. *Selectivity can be made* comparable to that of an expensive receiver by using a good active audio filter, such as that made by MFJ Enterprises, if you can afford the extra 8 mA current required. However, the backlash in the present tuning system makes it easy to lose the signal. A high-Q resonant transformer might be a good compromise.[9]

One disappointment with the original design was finding more than 100 mV rms of lo signal at TP1 in the product detector with the equipment running at 1.8 volts gate-to-gate lo voltage on Q1, Q2. This represents an unbalance voltage of unknown origin, which could be due either to unbalance in the transformers, phase unbalances, or even differing gains of each of the twin fets despite identical dc drain currents. Although back leakage through Q3 to the antenna must have been smaller than a few millivolts, such leakage could be detected as a signal comparable with low background noise, with the prototype receiver on the miniature antenna and a standard receiver on my big antenna. With two adjacent

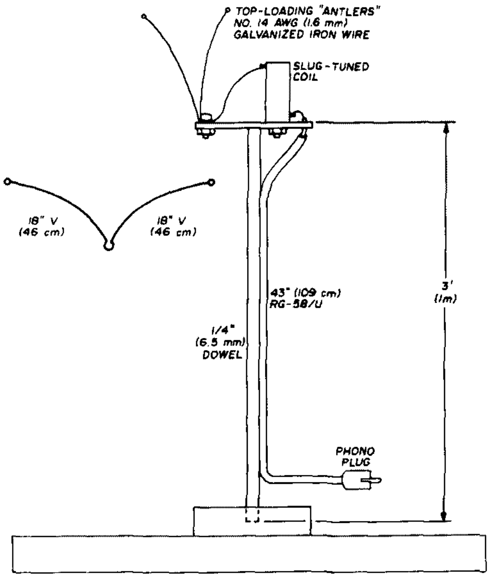



**fig. 10. Miniature resonant antenna for indoor use or as a portable outdoor antenna. Tuned circuit has a voltage gain of about 50, or 26 dB. At resonance, the antenna *has a* resistive impedance near 50 ohms.**

large antennas, the effect would be more serious, although the radiation would still be only microwatts. So I recommend, where amateur stations are close together, using the alternative product detector and adjusting R6 of fig. 2 for minimum antenna output, which should correspond to minimum lo unbalance voltage at TP1.

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**ham radio**

# the ground-plane antenna: its history and development

## Some useful information for those interested in the design and adjustment of this popular antenna

This article will be of value to those interested in designing and adjusting ground-plane antennas. The article describes the original invention of the ground-plane antenna, points out an error in assumptions regarding radiation resistance, and includes formulas for designing a simple matching network.

### background

A study of 30 - 60 MHz antennas made in 1936 by Dr. George H. Brown and J. Epstein of RCA brought to light two principal defects in most types of such antennas used at that time: the transmission line was not terminated properly, and sizable standing waves occurred on the outside of the coaxial transmission line. The J-antenna and the sleeve or coaxial antenna (both popular at that time) were found to be particularly susceptible to such standing waves.

Brown and Epstein found that the use of horizontal quarter-wavelength ground rods extending from the base of the vertical antenna established a virtual ground plane and shielded the coaxial line from the rf field of the antenna. Two types of coaxial matching networks were developed. Both supported the antenna rod mechanically, insulated it from ground at rf but grounded it at lightning and dc frequencies, and at the same time provided a good impedance match between antenna and transmission line at the operating frequency. Two patents resulted from this work: the original[1] was issued in 1941, and another was issued on an improved design[2] in 1942.

The original design, fig. 1, was built and tested in 1938 in Camden, New Jersey. It was my privilege to witness some of these tests. The first design used a quarter-wavelength antenna rod, four quarter-wavelength ground rods, and a quarter-wavelength coaxial support stub shorted at the bottom end. The coaxial transmission line was matched to the antenna base resistance (the reactive component was negligible when the antenna was exactly one-quarter wavelength long) by a quarter-wavelength coaxial line connected between the antenna base and the antenna end of the transmission line.

This quarter-wavelength line had a characteristic impedance of

$$Z = \sqrt{R_a Z_l} \qquad (1)$$

where

$Z$ = characteristic impedance of the quarter-wavelength matching section (ohms)
$R_a$ = antenna resistance (ohms)
$Z_l$ = transmission-line characteristic impedance (ohms)

This coaxial matching section can be compared to the so-called Q-section used by many amateurs years ago as an impedance transformer between an open-wire transmission line and a center-fed antenna.

### improved design

In the improved ground-plane design, fig. 2, the coaxial Q-section was eliminated, the transmission line was connected directly to the antenna base, the antenna was shortened to present an impedance of 19-j29.5 ohms, and the previous quarter-wavelength support section was shortened to about one-sixth wavelength. This shorted stub then had an inductive reactance of about +42 ohms which, in parallel with the capacitive antenna impedance of 19-j29.5 ohms, resulted in a parallel impedance of 65 ohms — the characteristic impedance of the coaxial transmission line used at that time. Each of the four horizontal ground rods remained a quarter wavelength long. Tests showed that their length was not too critical.

The values above were calculated as follows, as stated in Dr. Brown's article in *Electronics*:[3]

**By Harold C. Vance, Sr., K2FF.** (Mr. Vance became a silent key in August, 1976.)

The parallel inductive reactance of the matching stub required for parallel resonance with the capacitive reactance of the antenna is

$$X_s = \frac{R_a^2 + X_a^2}{X_a} \qquad (2)$$

where

$X_s$ = stub reactance (ohms)
$R_a$ = antenna resistance (ohms)
$X_a$ = antenna reactance (ohms)

The parallel impedance of the antenna, a pure resistance at resonance, is

$$R_p = \frac{R_a^2 + X_a^2}{R_a} \qquad (3)$$

where

$R_p$ = terminating resistance presented to the transmission line at the antenna base.

In his *Electronics* monograph[3] Dr. Brown points out that, ". . .it is practically possible to present a resistance that will match a concentric transmission line of any characteristic impedance above 25 ohms. . .but the antenna length becomes extremely critical when resistances in excess of 100 ohms are desired."

The evolution of the ground-plane antenna design is described in detail in an article by Dr. Brown and J. Epstein in the July, 1940, issue of *Communications* magazine.[4] This article also shows a ground-plane antenna with a close-spaced reflector. A gain of approximately 3 dB was obtained. The antenna and matching-stub lengths were changed slightly to maintain a match between antenna and transmission line when the reflector was added.

## erroneous assumption

In an article on the ground-plane antenna, Stephens[5]

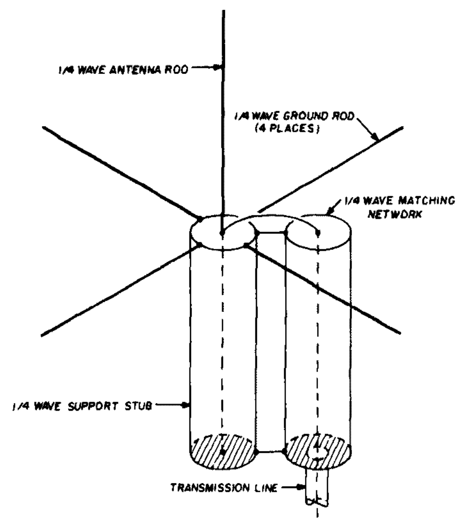



**fig. 1. Original ground-plane antenna design, which was built and tested by RCA in Camden, New Jersey in 1938. The coax matching section may be compared to the familiar Q-section transformer used by amateurs years ago to match open-wire transmission lines to center-fed wire antennas.**

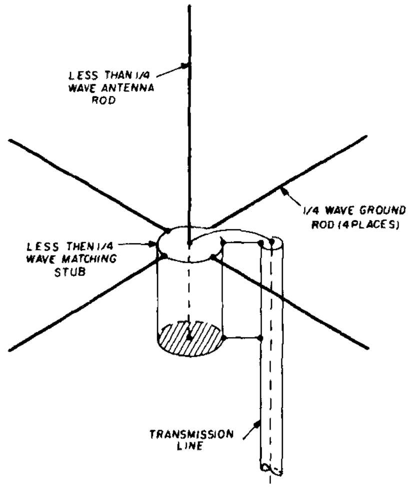



**fig. 2. Improved ground-plane antenna. The coaxial Q-section was eliminated, and other innovations resulted in an antenna that was easier to match and adjust.**

refers to an article by Hassenbeck[6] giving formulas and curves that claim to be values of resistance and reactance of a vertical antenna supported above four one-quarter wavelength ground rods. Hassenbeck states that data given by King and Blake[7] was used in establishing his curves.

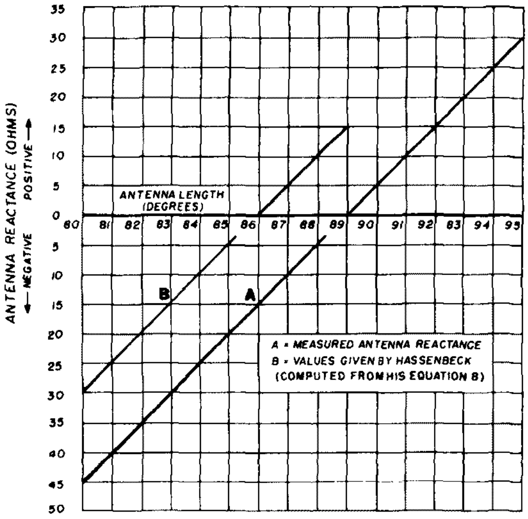
Dr. Brown pointed out[3] that, "the data given by King and Blake apply to a symmetrical antenna fed at its center or to a vertical antenna operating against an infinite metal sheet. . .there is a fundamental difference in the impedance of an antenna operating with four ground rods and an antenna operating over a semi-infinite sheet."

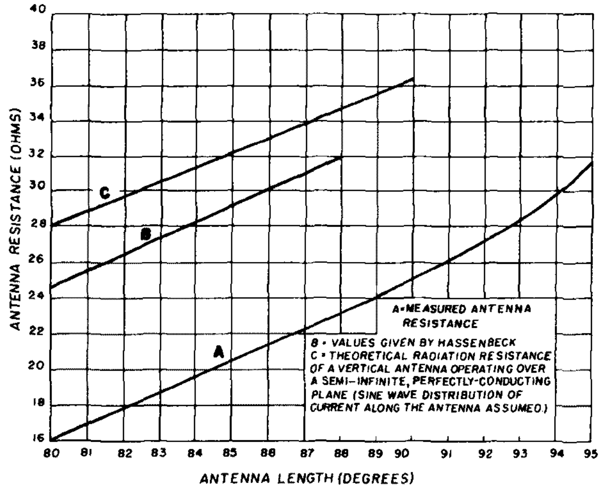
Actual measurements made by Dr. Brown showed that an antenna using four ground rods "has an appreciably lower radiation resistance than is generally assumed for the same antenna operating over a semi-infinite conducting plane." He stated that, "with an antenna exactly one-quarter wave in length, replacing the four ground rods by a metal disc one wavelength in diameter changed the actual measured radiation resistance from 25 ohms with the ground rods to 37 ohms with the one-wavelength-diameter disc."[3] Measurements of an earlier experimental *unit*[4] showed an antenna resistance of 21 ohms with ground rods.

The actual antenna resistance and reactance values shown in **fig. 3** were measured at 60 MHz. The ground rods were one-quarter wavelength at 60 MHz. The antenna-rod diameter was 0.625 inch (16mm), and its length was varied to those shown in the curves.

In Hassenbeck's article[6] the characteristic impedance of the support stub was assumed to be 70 ohms. In the RCA MI-7823-A antenna,[3] Dr. Brown used a stub having a characteristic impedance of only 41 ohms "to lengthen the support section when low frequencies requiring long antenna sections were used, and to make the adjustment of the shorting plug less critical."

In the case of the MI-7823-A antenna the length, $S$, of the inductive matching and support stub in electrical



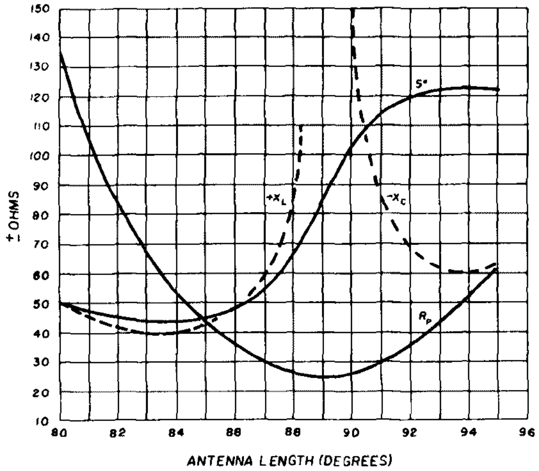




**fig. 3. Actual resistance and reactance values of the ground-plane antenna measured at 60 MHz compared with values given by Hassenbeck in reference 6 (top two graphs). Bottom set of curves shows characteristics of the RCA MI-7823-A ground-plane antenna including parallel-resonance resistance, $R_p$; shunt reactances, $X_L$ and $X_C$, required for resonance; and matching-stub length, $S^o$, required for the various values of reactance and resistance.**

degrees (360° = one wavelength) is given as:

$$tan\ S^\circ = \frac{X_{stub}}{41} \tag{4}$$

where $X_{stub}$ is the reactance of the matching stub (ohms) and $41$ is the characteristic impedance of the MI-7823-A matching stub (ohms).

In more general terms,

$$tan\ S^\circ = \frac{X_{stub}}{Z_{stub}}, \quad or\ S^\circ = arctan \frac{X_{stub}}{Z_{stub}} \tag{5}$$

where $Z_{stub}$ is the characteristic impedance of the stub (ohms).

That is, the stub length, $S$, in electrical degrees equals the angle whose tangent corresponds to the value obtained when $X_{stub}$ is divided by $Z_{stub}$.

## conclusion

The need for actual experimental measurements is well illustrated by the error of Hassenbeck's assumption that the radiation resistance of the ground plane antenna would be the same as that of an antenna operating over a semi-infinite sheet. Many factors such as proximity of other objects, incorrect assumptions, and structural differences can cause actual values to vary widely from calculated values. An R + jX bridge, such as those described in *QST*[8] and *ham radio*[9,10] will prove invaluable for this purpose.


## acknowledgement

I wish to thank Dr. Brown for furnishing me with copies of the referenced articles by him and J. Epstein.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**ham radio**

# broadband matching techniques for transistor rf amplifiers

The ins and outs of transistorized rf amplifiers can be effectively matched by using broadband matching techniques

**Ever since the introduction** of the transistor, there has been considerable interest in using the low-impedance characteristic of these devices to advantage in the design of broadband amplifiers. The most critical parts of a multi-octave amplifier are the input and output impedance transformation networks. In general, it is more difficult to accomplish large impedance transformation ratios, or to design networks which can transform low impedances. The latter because transmission-line transformers require a very low line impedance, which usually limits them to impedance ratios of less than 16:1 in 50-ohm systems; other types of transformers require tight coupling coefficients (or excessive leakage inductance will reduce the effective bandwidth).

Numerous methods have been successfully used in broadbanding vacuum-tube amplifiers. However, most of these methods *yield* relatively poor results when applied to low-impedance transistor circuits. Shunt-peaking, which works quite well with high-impedance tube circuits, becomes very difficult to use in transistor circuits because of the presence of $r_{bb'}$ and $r_{b'e}$ and the loading effect of $C_i$. In addition, the improvement obtained is significantly less than that realized for tube circuits.

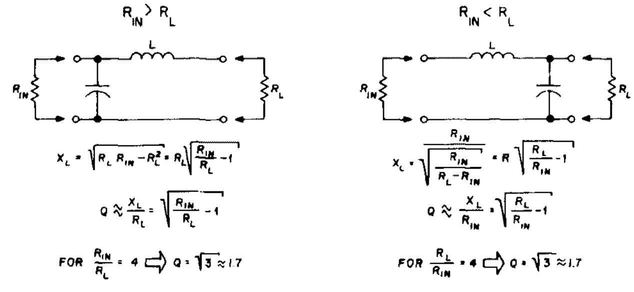



**fig. 1. L network calculations of Q.**

The L-network (**fig. 1**) which is a very simple network, suffers from two serious drawbacks with regard to broadband applications: first, the overall Q of the circuit is determined by the ratio of the impedances to be matched and *cannot* be set independently. Further, the higher the ratio of impedances to be matched, the higher the overall circuit Q. For a ratio of only 4:1, $Q = 1.7$, which is much too high for multi-octave use. The tapped-tuned circuit (**fig. 2**) allows the overall circuit Q to be set independently of the terminating impedances. However, there is a restriction on the minimum allowable value for Q which eliminates this circuit from consideration in multi-octave applications. The pi-network (**fig. 3**) suffers from essentially the same unacceptable restriction on minimum circuit Q as does the tapped-tuned circuit.

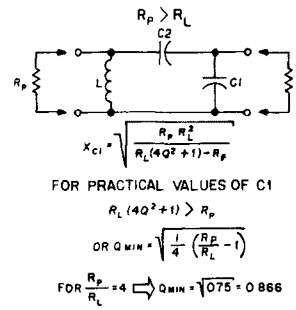



**fig. 2. Tapped-tuned circuit calculations of Q.**


**By Bill Leonard, WA7WHZ,** Technical Staff, Hughes Aircraft Company, 6573 Debs Avenue, Canoga Park, California 91307

## transmission-line transformers

The availability of high-frequency, high-permeability ferrites has prompted the development of wideband toroidal transformers which are ideally suited for multi-octave, low-impedance applications. There are numerous versions in use today, the more popular of which have been described in principle by Granberg.[1] Only the most popular of these, the transmission-line transformer, will be discussed here.

A wideband model of a toroidal 4:1 impedance-matching transformer was first investigated by Ruthroff,[2] and more recently by Pitzalis and Couse.[3] Transmission-line transformers yield nearly the best performance in terms of bandwidth and power handling capability. Units with bandwidth ratios of 20,000:1 (i.e., a Q of 0.007) and power capabilities in excess of 300 watts have been built. Figs. 4 and 5 show 1:1 and 4:1 impedance transformers. It is customary to show both the conventional and the transmission-line equivalent

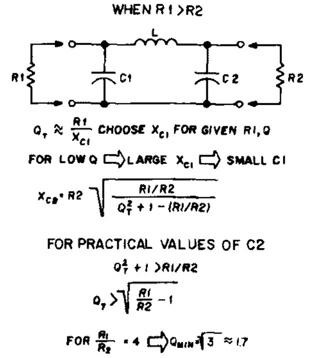



fig. 3. Pi-network calculations of Q.

circuits since the parameters which determine the lower cutoff frequency are very different from those which determine the upper cutoff frequency.

In conventional transformers, the interwinding capacitance resonates with the leakage inductance to produce a loss peak. This limits the high frequency response. In transmission-line transformers, the coils are arranged such that the interwinding capacitance is a distributed component of the characteristic impedance of the line. This characteristic makes the construction of low-impedance lines possible by use of close spacing (many twists per inch) of the wires which form the twisted pair. Although, in principle, any type of transmission line may be used, twisted pairs have the advantages of low cost and small size, while coaxial lines yield the best response due to their having relatively constant impedance as a function of length.

## frequency response

Analyzing the frequency response of broadband transformers is not a trivial exercise. The upper frequency limitations are related to the length of the transmission line and to the ratio between the characteristic impedance of the line and the load impedance. Usually, the length of

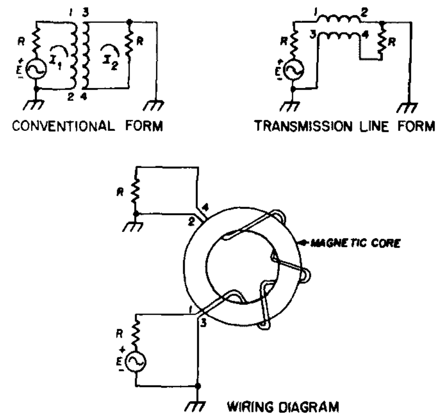



fig. 4. 1:1 impedance transformer shown in the conventional and transmission-line form along with the wiring diagram.

the transmission line is chosen to be λ/8 or less at the upper frequency desired. For a 4:1 impedance transformer, the characteristic line impedance ($Z_o$) should equal $\sqrt{R_L R_S}$. The shorter the physical length of the line, the less important exact matching becomes. The exact expression which describes the high-frequency behavior of the 4:1 impedance transformer can be found in references 2, 3, and 4. The result for matched impedances is

$$\frac{P_{in}}{P_{out}} = \frac{(1 + 3\cos\beta\ell)^2 + 4\sin^2\beta\ell}{4(1+\cos\beta\ell)^2} \qquad (1)$$

where $\beta$ = phase constant of the line $\left(\frac{2\pi}{\lambda}\right)$
$\ell$ = line length (same units as λ)
$\beta\ell$ measured in radians

Fig. 6 displays this expression as a function of line length. It can be seen that the upper cutoff (−3 dB) frequency occurs at a line length of 0.3λ for matched loads. However, in practice it is customary to keep $\ell < 0.125\lambda$ to minimize effects of mismatch.

For analysis of the lower cutoff frequency, the

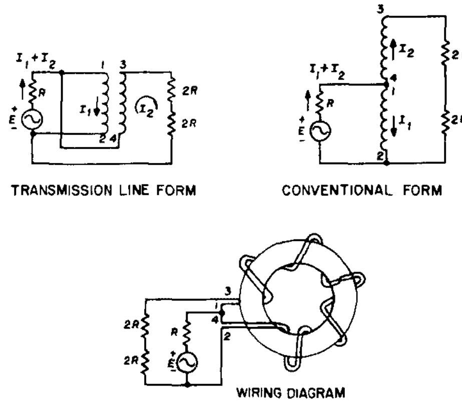



fig. 5. 4:1 impedance transformer shown in the same manner as the 1:1 transformer.

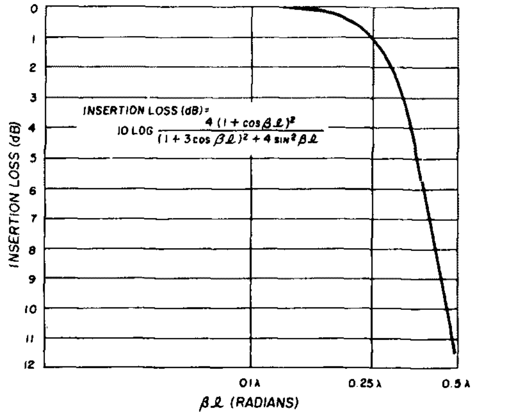



**fig. 6. Insertion loss versus line length for a matched broadband transformer.**

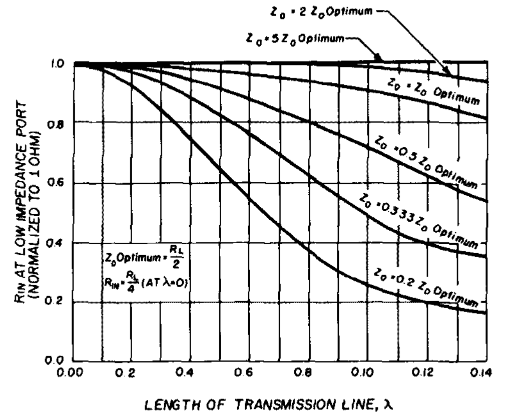



**fig. 7. The effect of various mismatched conditions shown as input impedance versus transmission-line length.**

conventional form of the equivalent circuit (fig. 5) is used. The lower cutoff frequency is determined by the falloff of primary reactance as the frequency is decreased. This reactance is determined by the series inductance of the transmission-line conductors. Therefore, the longer the line length, the greater the series inductance, and the lower the cutoff frequency. This is in direct conflict with extending the upper cutoff frequency which, as previously noted, is enhanced by shortening the physical length of the transmission line. Minimizing the overall length, consistent with achieving the necessary reactance for the low-frequency requirements, can be accomplished with high permeability materials such as ferrites. The inductance of a conductor is directly proportional to the relative permeability of the surrounding medium. A high permeability material placed close to the transmission line acts on the external fringe field in such a way as to significantly increase the effective inductance, thereby greatly lowering the lower cutoff frequency. The key point is that there is no influence upon the characteristic impedance of the line. Therefore, there is no degradation of the high-frequency cutoff characteristics. Also, power transferred from input to output is not coupled through the ferrite material, but rather through the dielectric medium separating the transmission-line conductors. This is an important characteristic since it allows for relatively small cross-section ferrite materials to be operated at very high power levels without danger of saturation. This is in contrast to the conventional transformer which couples power from primary to secondary entirely through the core which must be chosen to handle the total power without saturating.

In summary, the low-frequency cutoff requirements[3] can be related in the following expression

$$\ell \geqslant \frac{20R_L}{(1+\mu)f_L} \text{ inches} \tag{2}$$

Thus, the minimum allowable transmission line length can be specified in terms of the load resistance ($R_L$), the lower cutoff frequency ($f_L$ *in MHz*) and the relative permeability ($\mu$) of the core material.*

Impedance ratios in excess of 4:1 are easily obtained

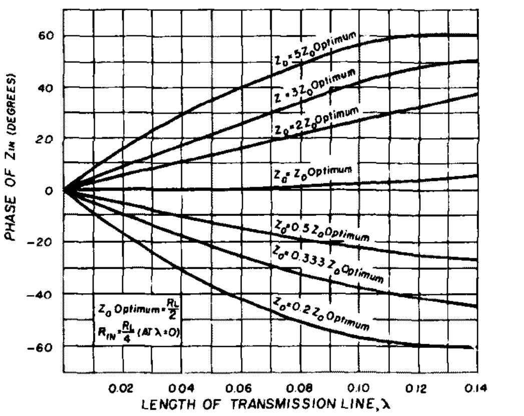



**fig. 8. The phase of input impedance, $Z_{in}$, versus transmission-line length for mismatched conditions.**

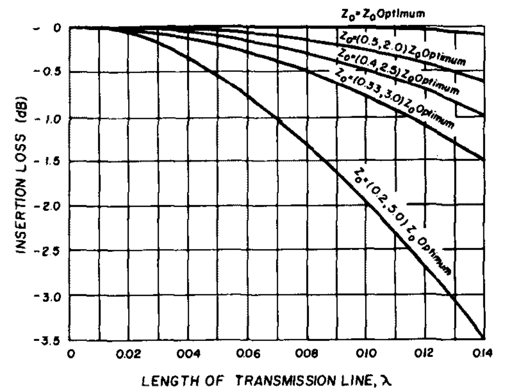
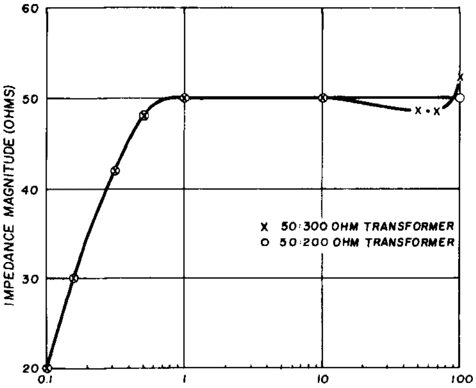
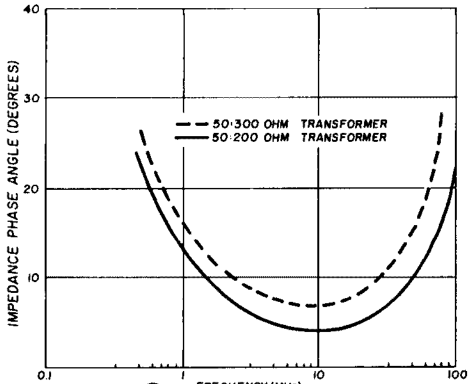


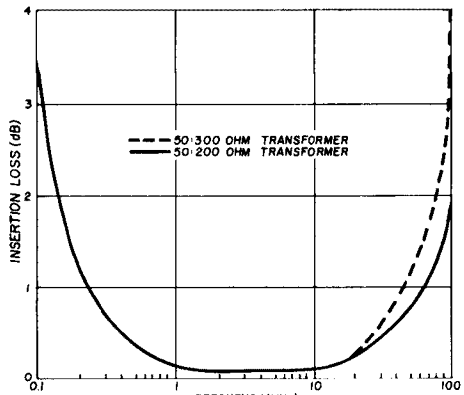

**fig. 9. The insertion loss versus line lengths for mismatched conditions.**



A IMPEDANCE MAGNITUDE



B IMPEDANCE PHASE ANGLE



C TRANSFORMER INSERTION LOSS VS FREQUENCY

**fig. 10. A comparision of two broadband matching transformers. Each transformer was wound on 2 Ferroxcube 266T125 3D3 cores. $A_T$ = 330±20% mH per 1000 turns, $\mu$ = 750 ±20%, $A_c$ = 0.183 cm$^2$. The winding for the 4:1 transformers is made from two no. 30 (0.25mm) vinyl-coated wires with 7 twists/inch. The length is 5.7″ (14.5cm) and 10 turns. The winding for the 6:1 transformer has a third 2.5″ no. 30 (6.5cm) wire added to form a three-wire transmission line. The transformers were designed for operation from 150 kHz to 100 MHz.**

by using the methods described in references 1, 3 and 4 when integer ratios (1:1, 4:1, 9:1,. . ., $n^2$:1) are acceptable. In addition, Krauss[4] discusses the construction of a three-wire transmission line for use in transformers which yields a nearly continuously variable (up to 6:1) ratio transformer. In general, characteristics such as bandwidth, passband ripple, and insertion loss began degrading as higher transformation ratios are required.

## mismatching

One question that immediately arises is the effect of a mismatch. Doherty and Hatley[5] have investigated this problem for a complex mismatch. Pitzalis and Couse[3] have resolved the question for a 4:1 impedance transformer encountering a resistive mismatch. These results are presented graphically in **figs. 7, 8** and **9**. These curves are normalized to $Z_o$ and ignore the low-frequency cutoff effects. They clearly demonstrate the relative insensitivity of the transformation with respect to a resistive mismatch for mismatches as high as 2:1. They also show that for $Z_o > Z_o$ (optimum), the input impedance is a series R-L circuit, while for the case where $Z_o < Z_o$ (optimum) the input impedance is that of a parallel R-C circuit. These facts can sometimes be used to advantage when compensation is necessary for stability or other reasons.

*To find the minimum line length in metric terms, use the following formula

$$\ell \geqslant \frac{51R_L}{(1+\mu)f_L} \quad cm$$

In conclusion, a comparision of two representative impedance transformers, as designed and tested by Krauss,[4] is presented in **fig. 10.** A conventional 4:1 impedance transformer is compared with a similar unit which uses a three-wire transmission line to obtain an impedance ratio of 6:1. It can be seen that, as was previously mentioned, the higher impedance ratio unit is slightly inferior in terms of overall performance.

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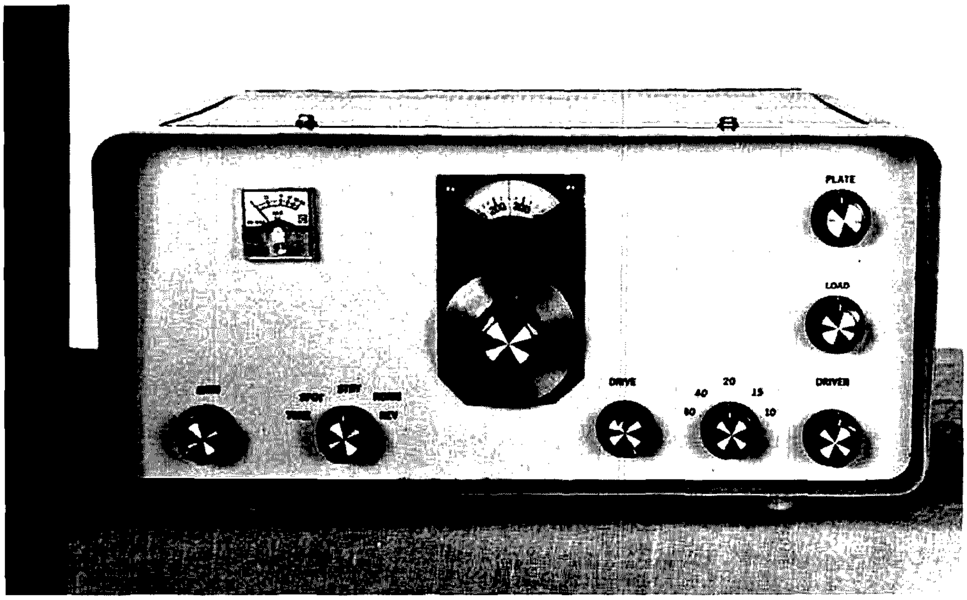



# 100-watt PEP five-band transmitter for CW and ssb

Except for driver and final this rig features solid-state stages, low cost, and easy construction

With the prospect of obtaining my general class license in the future, the need for an ssb transmitter arose along with the question of whether to build or buy. Considering the fact that a lot of experience and pleasure can be derived from homebrewing, I decided to build and came up with the following objectives:

1. Solid-state devices where practical.
2. 100 watts PEP.
3. Five-band coverage with CW and upper and lower sideband.
4. Low cost.
5. Easy construction.

After searching through various publications and not being able to find an article that fit all my requirements, I decided to design each stage separately using standard circuits and/or circuits from other articles.

By Harold Peters, WN3WTG, P. O. Box 1264, Tipton, Pennsylvania 16684

## description

A block diagram is shown in fig. 1. A 9-MHz dsb signal is generated with the speech amplifier and carrier oscillator, the carrier is removed by the balanced modulator, and the filter then removes one of the sidebands. The vfo and high-frequency oscillator (hfo) signals are mixed to produce the signal needed to heterodyne the 9-MHz ssb signal to the desired amateur band. This method results in less bandswitching than most other schemes.

The only disadvantage to this mixing scheme is that a sideband inversion occurs on 40 meters, and a tuning inversion occurs on 15 meters. On 40 meters the upper and lower sideband selection is reversed, and on 15 the low end of the band starts at the high end of the dial. But this isn't a major problem and one easily becomes accustomed to it.

The 9-MHz ssb signal and the output of the first mixer are mixed in the second mixer. The driver then amplifies this signal, and the Class AB final increases the signal level to 100 watts PEP output. Tubes are used in the driver and final to keep the cost down.

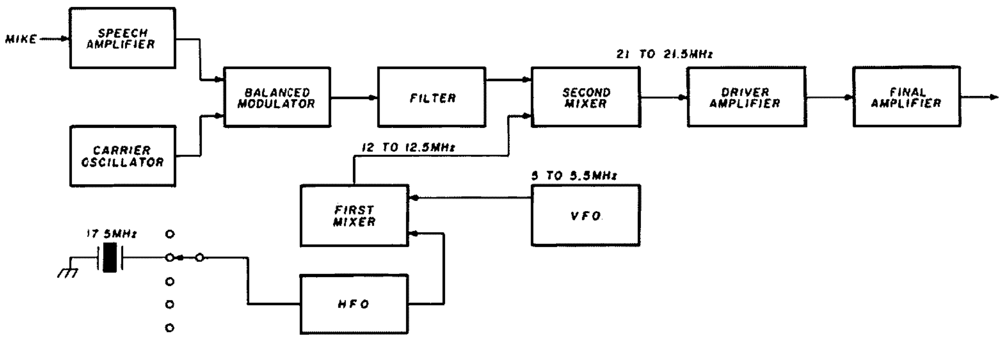
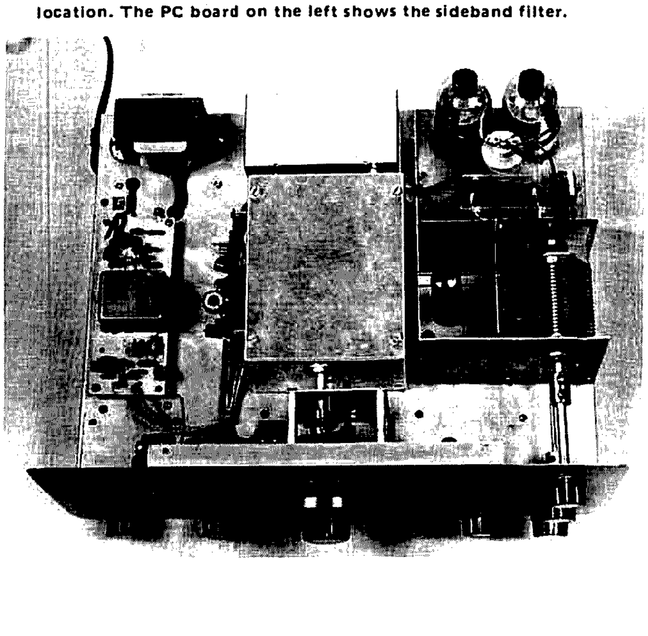



**fig. 1. Block diagram showing the five-band transmitter in the 15-meter position.**

**Top view of transmitter with cover removed to show component location. The PC board on the left shows the sideband filter.**



The entire transmitter was built a stage at a time and checked for proper operation before going on to the next stage. Each stage represented in the block diagram was built on a separate PC board or in the case of the vacuum-tube stages, in a separate compartment. Besides making construction and testing easier, this method allows for easy removal of a stage should it become defective.

## speech amplifier and carrier oscillator

The speech amplifier (fig. 2) consists of Q1, U1, and Q2.[1,2] The MPF102 provides a high-impedance input, and the 741 op amp and the 2N3053 provide ample audio power to drive the balanced modulator.

The speech amplifier as well as the carrier oscillator, high-frequency oscillator, and first and second mixer stages were constructed on double-sided glass epoxy PC board. The components were mounted on one side of the board, and the copper was left intact on the reverse side to act as a ground plane for added circuit stability.

The carrier oscillator (fig. 2) uses diode switching for selection of upper or lower sideband.[1,3] The trimmers for the crystals vary the carrier frequency to the desired location on the 9-MHz crystal filter skirt.

The operation of the carrier oscillator was verified by listening for the signal in a general-coverage receiver. The speech amplifier was tested by applying an audio signal to the microphone input and listening for the output with headphones.

## balanced modulator and filter

The outputs of the speech amplifier and carrier oscillator are mixed in the MC1496 balanced modulator[2] and

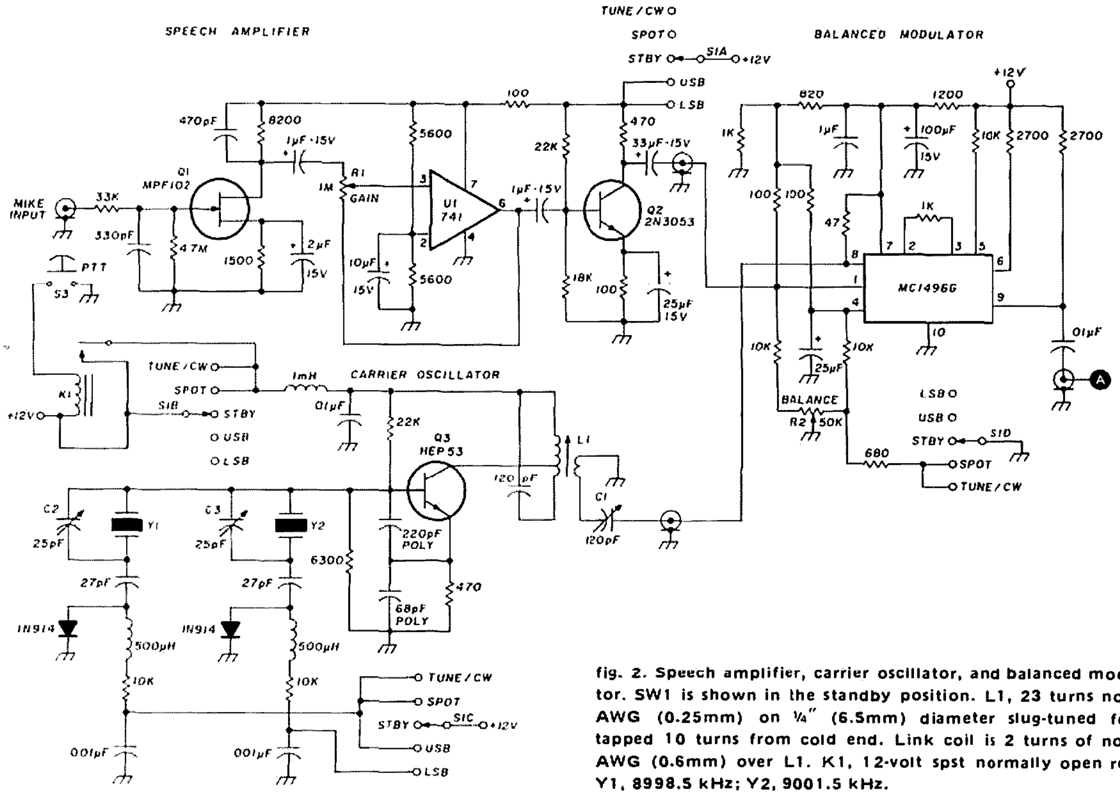



fig. 2. Speech amplifier, carrier oscillator, and balanced modulator. SW1 is shown in the standby position. L1, 23 turns no. 30 AWG (0.25mm) on ¼" (6.5mm) diameter slug-tuned form, tapped 10 turns from cold end. Link coil is 2 turns of no. 22 AWG (0.6mm) over L1. K1, 12-volt spst normally open relay. Y1, 8998.5 kHz; Y2, 9001.5 kHz.

the carrier is suppressed (fig. 2). In the CW mode, balance is upset and the carrier is allowed to pass through.

The 2N2222 (Q5) of the filter stage (fig. 3) acts as a buffer and also serves as the keying point for CW by applying keyed +12 volts from Q6.[4,5] Q7, a 40841 gate-protected dual-gate mosfet, makes up for filter signal loss and acts as the ALC control point.[1]

These stages were constructed on single-sided glass epoxy PC board. Care was taken during construction to ensure that no signal leak through would occur to destroy the carrier and opposite sideband suppression.

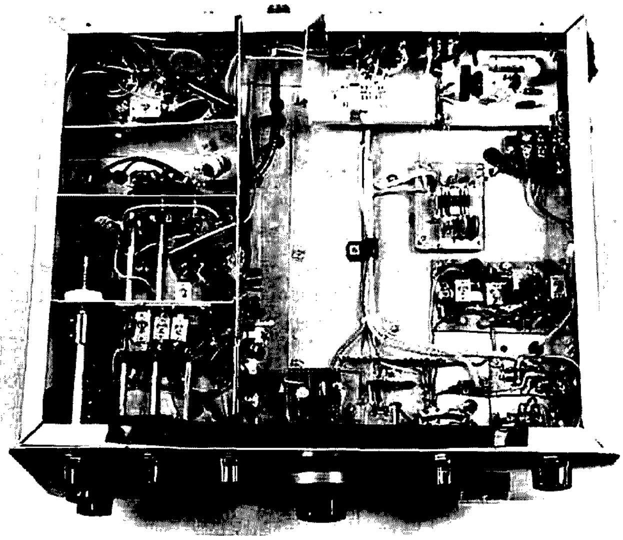
With the previous stages operating properly, the output of Q7 is a 9-MHz ssb signal, which can be monitored on a receiver. L1, C1, and R2 (fig. 2) are adjusted for maximum carrier suppression and C2, C3, C4, and C5 are adjusted for best voice quality consistent with good carrier and opposite sideband suppression. L2 is adjusted for maximum output.

## vfo and hfo

The schematic for these stages is shown in fig. 4. The vfo, Q8, is a series-tuned Clapp circuit and includes a two-stage buffer amplifier, Q9, Q10, and a lowpass filter for a very stable and clean output.[1,6] Voltage to the MPF102 is regulated by a 9-volt zener.

The vfo PC board was mounted in its own enclosure with angle brackets, and the box was mounted in the center of the chassis. Phono jacks couple power and vfo output to the enclosure to make for easy removal of the vfo for testing and temperature compensation. Using a receiver or frequency counter, the vfo was adjusted to

Bottom view of the transmitter. The audio and modulator stages are located in the lower-right corner. The power regulator is in the upper right.



tune from 5-5.5 MHz by adjusting the bandset trimmer and L9.

The hfo was also mounted in a separate enclosure. The crystals are diode switched so that the hfo could be mounted away from the bandswitch.[3] This stage is not used on 80 or 20 meters, as the vfo 5-MHz output is mixed with the 9-MHz ssb signal for output on these bands.

placed on the bandswitch, and the mixer PC boards were placed close by to keep lead lengths short.

With the aid of the grid-dip meter, the toroid/trimmer assemblies were checked and adjusted for approximate frequencies before placing them in the circuit.

## driver and final amplifier

The driver stage, shown in fig. 3, uses a 6GK6 power

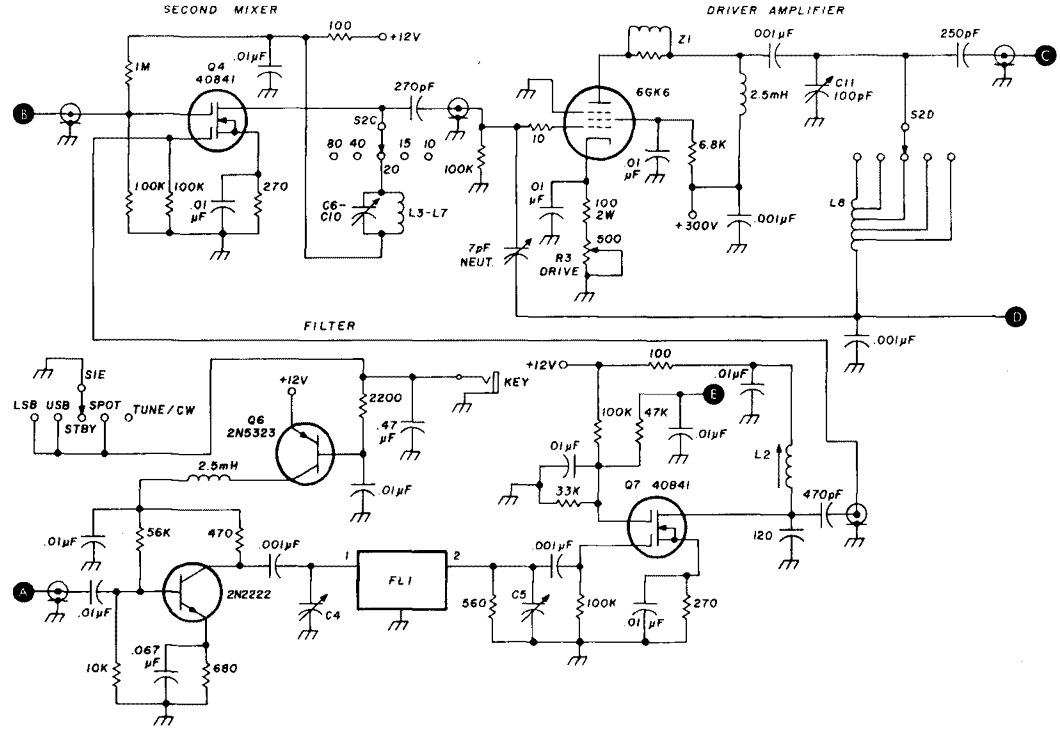



**fig. 3. Second mixer, driver, and filter. FL1, 9-MHz crystal filter, KVG XF-9A. C4, C5, 35-pF trimmer. Z1, 4 turns no. 20 AWG (0.8mm) on 47-ohm, 1-watt resistor. L2, 20 turns no. 30 AWG (0.25mm) on T-50-6 form; C6, 250 pF. L4 (40 meters), 22 turns no. 28 AWG (0.3mm) on T-50-6 form; C7, 250 pF. L5 (20 meters), 14 turns no. 22 AWG (0.6mm) on T-50-2 form; C8, 120 pF. L6 (15 meters), 12 turns no. 22 AWG (0.6mm) on T-50-2 form; C9, 120 pF. L7 (10 meters), 9 turns no. 22 AWG (0.6mm) on T-50-2 form; C10, 120 pF. L8, 60 turns no. 28 AWG (0.3mm) on T-68-2 form, tapped at 30, 12, 6, and 3 turns for 40-10 meters respectively.**

## first and second mixers

The first mixer[7] uses a 40841 to mix the vfo and hfo outputs. The proper tank circuit is switched by the bandswitch, with the same tank used for both 80 and 20 meters. The output frequencies of the first mixer are: 5-5.5 MHz for 80 meters, 16-16.5 MHz for 40 meters, 5-5.5 MHz for 20 meters, 12-12.5 MHz for 15 meters, and 19-19.5 MHz for 10 meters.

Q4, another 40841 (fig. 3), serves as the second mixer and uses a circuit similar to that of the first mixer. Second-mixer output is a low-power version of the ssb signal to be transmitted.

All coils in the mixer tanks are wound on toroids for compactness and self-shielding. The windings were spaced on the toroids to occupy the entire circumference of the core. Each toroid and its trimmer were

pentode tube.[8] A drive control is included to vary the amount of drive to the final. The driver tank coil differs from the previous tank coils in that it is was wound on a single toroid, tapped for each band, and tuned to resonance with C11 from the front panel of the transmitter.

The final stage (fig. 5) uses two 6146Bs in a class-AB parallel configuration. Plate voltage is supplied through RFC1, a Johnson 102-572 transmitting-type rf choke. Meter M1 measures plate current.

The ALC voltage is obtained from the grid bias. It is rectified and applied to the filter stage through R4, which controls the amount of ALC voltage. The ALC circuit was adjusted with the aid of an oscilloscope. A two-tone test signal was injected into the microphone input, and the transmitter rf output was monitored on

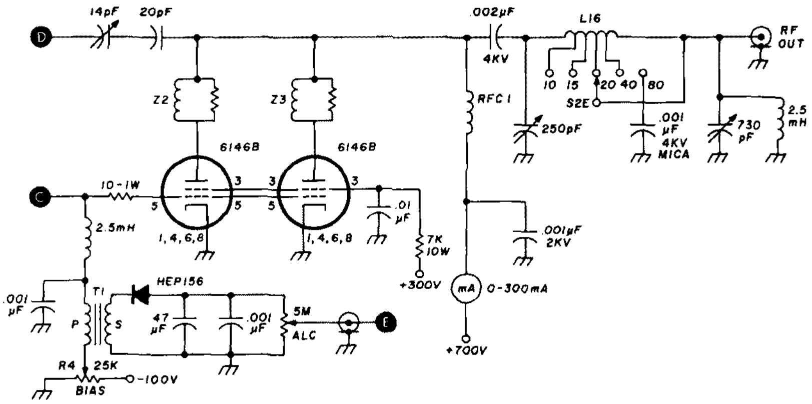



fig. 5. Final amplifier. T1, audio transformer; primary 600 ohms, secondary 2000 ohms. Z2, Z3, 5 turns no. 20 AWG (0.8mm) on a 47-ohm 1-watt resistor. L16, 38 turns no. 12 AWG (2.1mm) on T-200-2 form, tapped at 4, 6, 11, and 22 turns for 10-40 meters respectively.

the scope. Audio gain control R1 was advanced until flat-topping just started to occur; then R4 was adjusted to eliminate this distortion.

Conventional procedures were used in adjusting the neutralizing and the final plate and loading capacitors. Plate current at maximum output is about 220 mA.

## 12-volt power supply

Fig. 6 shows the regulated 12-volt power supply, which was included on the chassis so that during non operating periods the power to the vfo can be left on. A 723 IC regulator and a 2N3772 pass transistor provide about 3 amps at 12 volts.[9] R3 sets the output voltage.

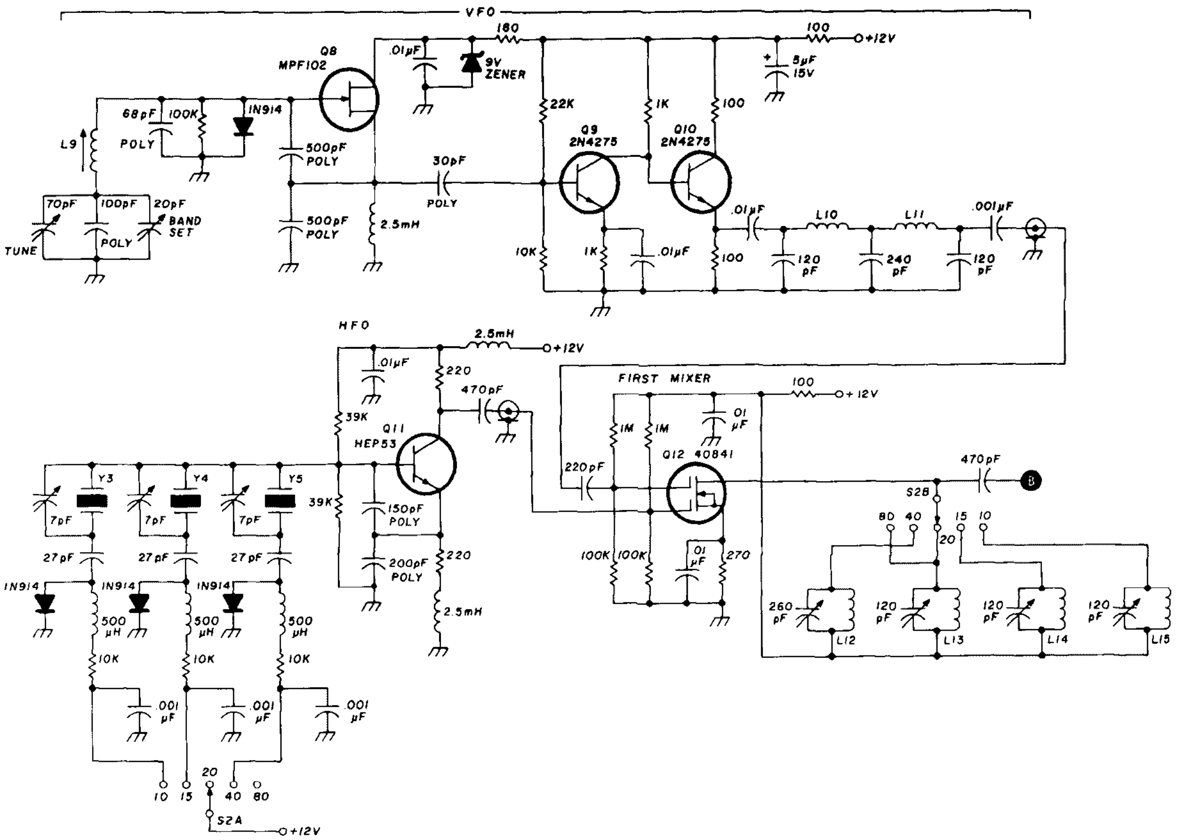



fig. 4. Vfo, hfo, and first mixer. S/2 is shown in the 20-meter position. L9, 35 turns no. 32 AWG (0.2mm) on ¼" (6.5mm) slug-tuned form. L10, L11, 38 turns no. 32 AWG (0.2mm) on T-37-2 form. L12, 12 turns no. 22 AWG (0.6mm) on T-37-2 form. L12, 31 turns no. 28 AWG (0.3mm) on T-50-6 form. L14, 16 turns no. 22 AWG (0.6mm) on T-50-2 form. L15, 12 turns no. 22 AWG (0.6mm) on T-50-2 form. Y3, 14 MHz. Y4, 17.5 MHz. Y5, 11 MHz.

The 2N3772 was mounted on the back of the chassis, which acts as a heat sink.

## closing remarks

A small change to be made in the future is to place the spot function on a separate front-panel switch. As it

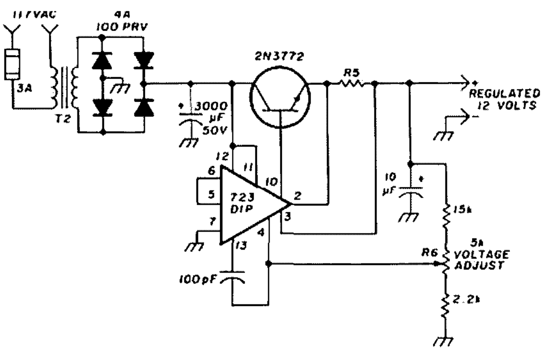



**fig. 6. Regulated power supply for the solid-state stages of the transmitter. T2, 24-volt 2-amp power transformer. R5, 0.1-ohm resistor consisting of 8 feet (2.4m) of no. 22 AWG (0.6mm) wire wound on a wooden dowel.**

is now, the spot function works only in the upper sideband position. But the opposite sideband can be heard in my receiver because of the close proximity, so it poses no real problem.

## acknowledgements


I would like to thank all of the authors whose ideas and circuits I borrowed for my transmitter. Special thanks go to Howard Stark, WA4MTH, for his encouragement and assistance.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# automatic 600-kHz up/down repeater-mode circuit for two-meter synthesizers

## Although designed for use with the GLB 400B Channelizer, this circuit may be used with other synthesizers using BCD coding to their divide-by-*n* counters

With the continuing increase in the number of repeaters appearing on two meters, most amateurs are finding that being "rock bound" leaves a lot to be desired, not to mention the cost of filling those crystal sockets in most two meter rigs. The trend seems to be that more amateurs are finding that synthesizers allow them to work the entire band, and, in most cases, that they can work it in 5-kHz increments.

Some synthesizers on the market today use a pair of ganged switches, one to set the repeater input, the other the output frequency. This arrangement doesn't pose a problem except when it comes to mobile operation. Changing repeater frequencies while "threading the needle" in rush hour traffic can be exciting, if not downright hazardous.

This article deals with my answer to improving the ease of operation of the GLB 400B Channelizer, but the circuit may be used with other synthesizers using BCD coding to their *divide-by-n* counters.

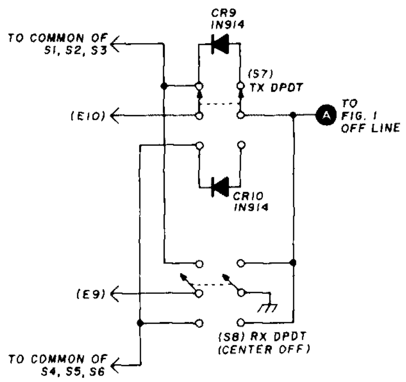


**fig. 1. Wiring diagram of the replacement switches for S7 (transmit select) and S8 (receive select). Rear view is shown.**

### features

The addition of the circuit of **fig. 2** provides an automatic 600-kHz *up* or *down* shift of the receive frequency and the ability to work repeaters on just one


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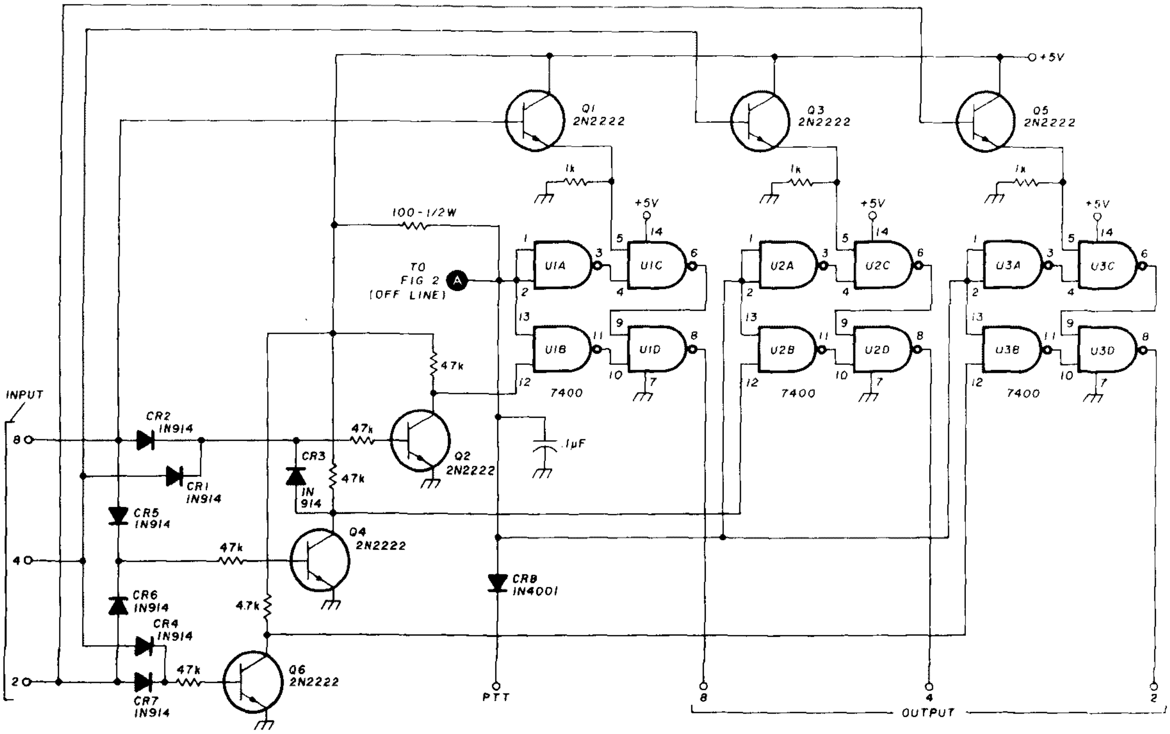



**fig. 2. Schematic of the automatic 600-kHz up/down repeater-mode circuit. The 100-ohm resistor is ½ watt; all others are ¼ watt.**

control head. The other set of switches can be left on a favorite repeater or simplex frequency and monitored by a simple flick of the transmitter select switch.

This circuit will work with all standard repeater pairings of 146.01 to 146.37 MHz and 147.60 to 147.99 MHz input and will shift to the simplex mode when the control switches are set to the standard simplex frequencies of 146.40 to 146.59 MHz and 147.40 to 147.59 MHz.

Upon setting the repeater input in the control head, such as 146.28 MHz, the circuit changes the BCD input to the 100-kHz divide-by-*n* counter from a 2 to an 8, which offsets the receiver 600 kHz *up* and allows the receiver to operate at 146.88 MHz. Setting the repeater output frequency in the control head allows paralleling the repeater by offsetting the receiver 600 kHz *down*. In either case, the transmitter will always be on the frequency that appears on the control switches when the microphone PTT switch is depressed. Repeaters in the 147.00-MHz segment, such as 147.63 MHz in and 147.03 MHz out, are working in the same manner.

The combination of transistors Q2, Q4 and Q6 and their associated diodes provides the logic to the circuit. ICs U1, U2 and U3 act as three separate spdt switches that place the logic function in or out and are controlled by the *center off* position of the receive select switch (S8). Q1, Q3 and Q5 act to interface the ICs.

## construction

A full-size PC board layout for the automatic up/down repeater mode circuit is shown in **fig. 3**; **fig. 4** shows component location. Price information on etched and drilled PC boards may be obtained by sending a self-addressed stamped envelope to the author; however, the circuit is not critical and can be built on Vector board if desired. I would advise against substituting

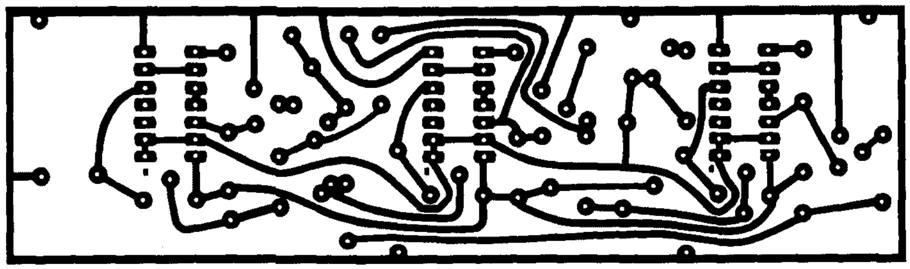


**fig. 3. Full-size PC-board layout for the automatic 600-kHz up/down repeater-mode circuit. Send a self-addressed stamped envelope to the author for prices on etched and drilled boards.**

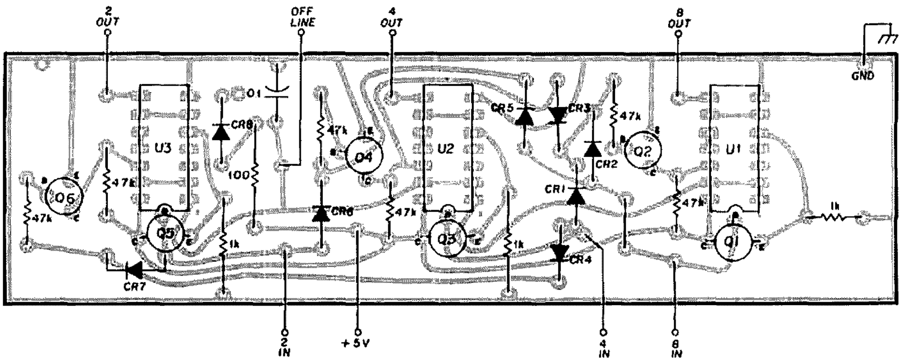


fig. 4. Reverse side of PC board showing parts location. CR8 is a type 1N4001; all other diodes are 1N914s. Use a ½-watt resistor for the 100-ohm unit; all others may be ¼ watt.

components or resistance values. The use of IC sockets is a must when using surplus ICs.

The Channelizer requires no outward change to its appearance, and only the transmit select switch (S7) need be changed from an spdt to a dpdt. The receive select switch (S8) is changed from an spdt to a dpdt (center off).

Rewire the new switches, S7 and S8, as shown in fig. 1. Diode CR9 and CR10 mount on S7. The ground shown on S8 and the 600 kHz offset should be connected to the ground of the Channelizer main board. The "off line" connects to the 600-kHz offset circuit board.

## checkout

The circuit should be checked out before installation as follows. Connect 5 volts to the $V_{cc}$ point and apply a ground to the circuit. With a voltmeter connected to pin 8 of U2, an indication of 4 or more volts should occur. U3 pin 8 should indicate the same positive voltage. U1 pin 8 should read less than 1 volt.

Now connect 5 volts through a 1k resistor to the BCD 2 input and note that pin 8 of U1 is now at 4 or more volts, while pin 8 of U2 and U3 are less than 1 volt. With this connection to the 2 input, place 5 volts through another 1k resistor to the BCD 4 input. Now pin 8 of all ICs should indicate less than 1 volt. Disconnect the resistors and touch one of them to the BCD 8 input. Pin 8 of U1 and U2 should be less than 1 volt, while U3 pin 8 should be 4 volts or more. Place a ground connection to the *off* line and note that when the voltage is again applied to the inputs (one at a time), the corresponding outputs will be high, while the others remain less than 1 volt.

## installation

The board is connected to the synthesizer by grounding it to the main board of the Channelizer and applying 5 volts to the $V_{cc}$ point from E5 or some other convenient source. Connect the PTT switch to E8 of the main board. Connect the *off* line to the switch as shown in fig. 1. Cut the wire from the junction of CR15 and CR16 to E20 and connect the end nearest the diodes to the BCD 8 input. Connect the other wire, nearest E20, to the BCD 8 output. Likewise, cut the wire going to E19 and connect to the BCD 4 input and output. Cut the wire to E18 and connect to BCD 2 input and output (see fig. 5).

## operation

Selecting the center position of the receive switch (S8) on the front of the Channelizer enables the 600-kHz repeater mode circuit. Changing the transmit switch (S7) to either up or down position selects the control head to be used, and, as mentioned earlier, two

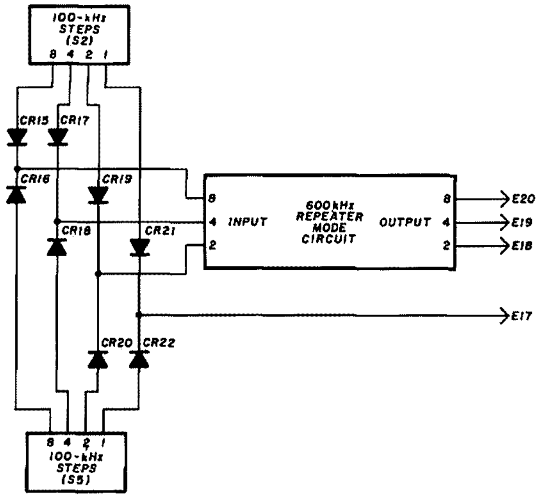


fig. 5. Diagram for connecting the 600-kHz repeater-mode circuit to the GLB 400B Channelizer 2-meter synthesizer (see text).

different repeater frequencies may be set up. But always remember that the repeater input frequency should be used for working *into* the repeater. Selection of the standard simplex frequencies will prevent the circuit from shifting the receiver, so simplex operation is provided, although the repeater mode is still selected by the center position of the receive switch (S8). Setting this switch to either of its extreme positions will disable the 600-kHz *up* or *down* circuit and return the synthesizer to its normal operation.

**ham radio**

# simple diode tester

## An autotransformer controls input voltage to this useful addition to your test equipment

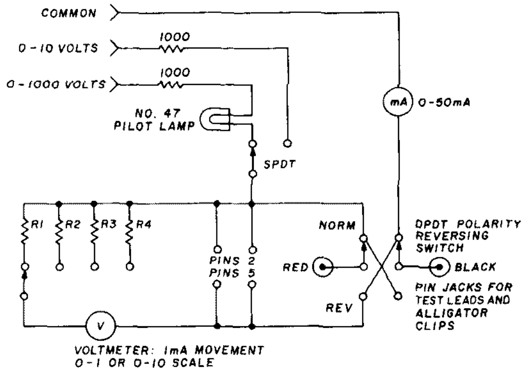
When you have a diode with no numbers on it and no color bands to identify it, what do you do with it? Here's a simple tester to help identify whether the diode is silicon or germanium. The diode current rating can be approximated by the physical size. Refer to the size and ratings from catalogs or other diodes you may have that are of known ratings. Small-signal diodes such as 1N914s and 1N34s will have a PIV rating somewhat above 25 volts. This tester will indicate that some small-signal diodes will have a lot of leakage current compared to the forward current, which is usually detrimental. It will also indicate zener voltage ratings.

### applications

Other uses for this tester include checking diacs, SCRs, triacs, LEDs, VR tubes, neon tubes, and pull-in and drop-out voltage of small relays, showing that many relays marked 24 volts will work nicely at 12-13 volts. (Multileaf relays may not be advisable at this lower voltage, however). You can form new or old electrolytic capacitors. After an electrolytic capacitor has been formed for a few minutes, the operating voltage for that particular capacitor may be safely marked at about 20 percent less voltage than the maximum forming voltage. That is a voltage where the capacitor is drawing something less than about 5 mA.

This tester was first built and used in 1942 and has been indispensable all these years. An earlier article[1] described this tester, but it was primarily designed to test VR tubes. When 1000-volt PIV diodes became available, it was necessary to modify the tester.

Modifying the tester to accommodate 1000 volts, that is, getting rid of the vacuum-tube dual triode (grid-controlled rectifier in this case) was easy since I




R1 = 1 meg = 1000 volts full scale
R2 = 100k = 100 volts full scale
R3 = 10k = 10 volts full scale
R4 = 1k = 1 volt full scale

**fig. 1. Metering and control section of the diode tester. Pins 2 and 5 are for a 7-pin miniature tube socket and an 8-pin octal socket for testing VR tubes. All resistors are 1 watt.**

**By Lloyd Jones, W6DOB,** 17779 Vierra Canyon Road, Salinas, California 93901

happened to have a couple of small Variac controls. With a Variac to control the input to the transformer primary and a solid-state half-wave rectifier with a 2 $\mu$F capacitor, it was easy to produce 1000 volts.

## description

If you buy and use any of the diodes (new or surplus), you should build one of these testers. The basic requirement for such a tester is a dc voltage supply that's variable from 0 to at least 500 volts, preferably to 1000 volts. This supply should be "soft"; that is, one with very poor voltage regulation. At 50 mA, the voltage as shown on the voltmeter may drop to 60 or 70 percent of the no-load voltage. (Most of the testing I do on this tester is at less than 5 or 10 mA). When I test unknown surplus diodes it's not uncommon to zap a few of the bad ones, so a no. 47 pilot lamp is connected in series with the 0-50 mA meter to protect the meter movement.

A second meter is used as a voltmeter. An ideal meter for this purpose is a 0-1 mA meter with series resistors adjusted to read 10, 100 and 1000 volts. In my tester another voltmeter with a pushbutton is used to read voltages less than one volt. Ideally, diode rectifiers should be added to the unused filament winding (6.3 V) when using the 1-volt meter to determine whether a diode is germanium (around 0.45 volt) or a silicon (around 0.7 volt).

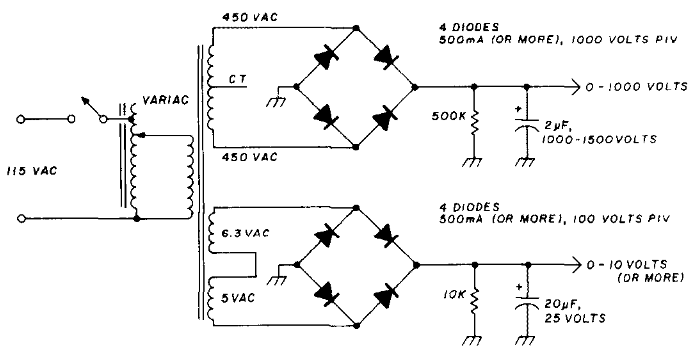



**fig. 2. Diode-tester power supply. An autotransformer is a must for input voltage control; so-called "light dimmer" controls and motor-speed controls will not control the ac input voltage.**

## circuit

Consider the tester in two sections. The first is the system following the rectified voltage; the second is for controlling the voltage from zero to some higher value of voltage. **Fig. 1** shows the schematic for the most versatile tester. This system uses a transformer and bridge-rectifier power supply (**fig. 2**) to supply 0-100 and 0-1000 Vdc. (By using the 6.3- and 5-volt windings in series and another bridge rectifier, you'll have more than ample voltage for 0-1 and 0-10 volts. Most small power transformers have these secondary windings).

A dpdt switch is used for quickly reversing the connections to the item under test, which eliminates changing clip leads. An spdt switch selects 0-1000 volts from the high-voltage supply or 1-10 volts from the

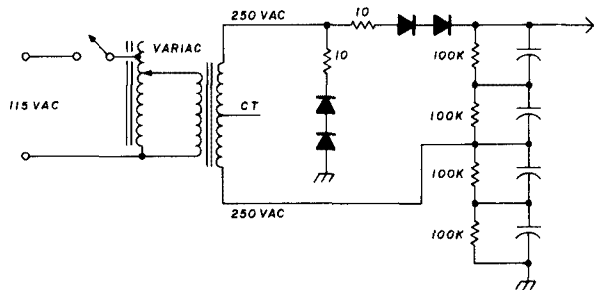



**fig. 3. Alternative power supply using a smaller transformer in a voltage-doubler circuit. Diodes are 1000 PIV, 500 mA (or more); all electrolytic capacitors are 40 $\mu$F, 450 volt; resistors are 1 watt. The 2 $\mu$F capacitor must be oil filled.**

low-voltage supply. A miniature 7-pin tube socket and an octal socket are used for testing VR tubes.

The power supply is shown in **fig. 2**. If you don't have a 450-volt center-tapped transformer, you can use the full-wave voltage-doubler in **fig. 3**.

Here's an important warning: while checking the PIV of a diode, *be sure to keep the current low*. Ten microamperes is fine; 100 microamperes is the maximum that can be used (for one or two seconds). One-hundred microamperes is just discernible on the 50-mA meter by the slightest movement of the pointer. The zener rating of any diode can be read directly on the voltmeter. My collection of zeners includes devices as low as 3 volts and as high as 175 volts. Some are soft; most are hard or stiff, meaning they produce good regulation.

## closing remarks

By this time it may have occured to you to keep all your old-fashioned parts. Some you can use; some you can give away to other experimenters. The days when you could walk into a radio parts house and buy amateur components are a thing of the past. However, you can find good bargains in surplus components by watching the ads in the amateur magazines. I bought a pound (half a kilogram) of diodes from M. Weinschenker (*ham radio*, May, 1973, page 60) for $10.00 and found that nearly 85 percent were good. About 100 were 1 amp, 1000 PIV; perhaps 100 were small zeners; another 100 or so were small-signal diodes including some noise diodes.

## reference

1. Lloyd Jones, W6DOB, "Tester for Glow Tubes, Selenium Rectifiers, and Germanium Rectifiers," *CQ*, November, 1957, page 130.

**ham radio**

# Q measurement and more

## With a communications receiver, this simple noise generator allows you to determine the loaded Q of tuned circuits

**This article describes** a simple piece of test equipment that can often replace a signal generator. When used with a communications receiver it can outperform the usual Q meter by measuring the Q of a tuned circuit while in place in the equipment. The principle, borrowed from the antenna noise bridge, is to energize the tuned circuit with noise, giving constant signal strength over a wide frequency band. The frequency response is then observed using a communications receiver. It is of course necessary that the receiver bandwidth be appreciably narrower than that of the circuit under test, but this is generally true of most communications receivers. The receiver S-meter is used to find the −6 dB bandwidth of the circuit, and from this the Q is calculated.

### the noise generator

The schematic is shown in **fig. 1**. Noise is generated by reverse breakdown of the base-emitter junction of Q1. Q1 output is then amplified by the remaining three transistors. The output, which is at low impedance, is S9+26 dB up to 7 MHz and S9 at 28 MHz, which allows very loose coupling to the circuit under test. Several transistors should be tried for Q1 and the noisiest one used. Note that, because the zener voltage is usually about 6 or 7 volts, the zener current and hence the noise output is very sensitive to battery voltage. A good battery is therefore necessary for stable output. Leads must be short to avoid resonance. The components are mounted, complete with a 9-volt transistor battery, in a metal enclosure measuring 3x4½x1 inch (77x108x25.5mm). A shield partition is installed across the enclosure (dashed line, **fig. 1**), and the electronic components are mounted on a terminal strip with ground connections soldered directly to the enclosure. This construction ensures good shielding, which is essential if the generator is to be used for receiver sensitivity measurements. Note the feedthrough capacitor in the power lead. The on/off switch is arranged to provide the alternatives of an external 12-Vdc supply through an "idiot diode" or 6.3-Vac operation. The 6.3 Vac modulates the noise output, making it easily distinguishable from other noise sources.

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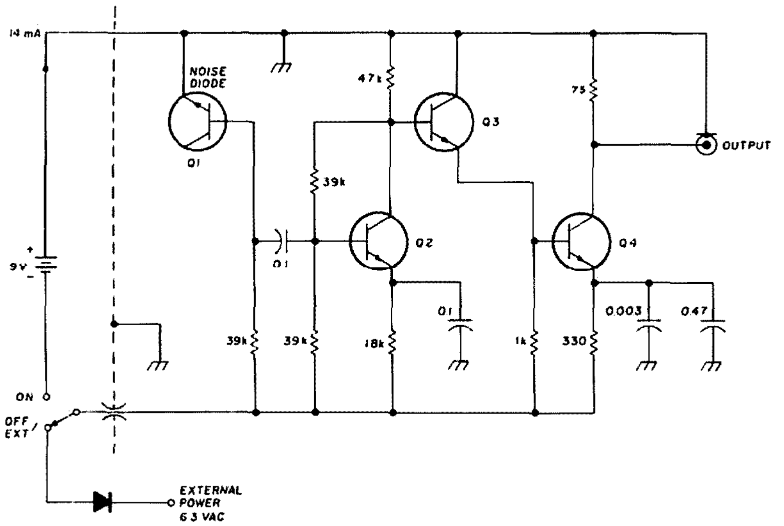



fig. 1. Noise generator schematic. Transistors are type 2N2368; resistors are 1/8 watt; capacitors are ceramic. Noise output is sensitive to battery voltage; therefore a good battery is necessary for stable output. An external ac power supply may also be used (see text).

## Q measurement

Fig. 2 shows one test setup for Q measurement. Here, both generator and receiver are coupled by the smallest capacitors that give a usable signal at the receiver, which minimizes loading effects. Either or both couplings may

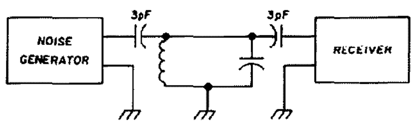



fig. 2. Example of a test arrangement for measuring Q. Minimum coupling between noise generator and receiver is required to minimize loading effects.

be inductive if convenient. The −6 dB point will be 1 or 2 units below maximum (there seem to be two rival standards; if in doubt, make a 6 dB attenuator and use it with the noise generator to check). Then, assuming sufficiently narrow receiver bandwidth, the formula to use is

$$Q = \frac{170\,(f_1)}{\%\ bandwidth\ (f_2 - f_3)} \tag{1}$$

where

$f_1$ = peak response (MHz)
$f_2$ = upper 6-dB point (MHz)
$f_3$ = lower 6-dB point (MHz)

For example, if the peak response is at 14.27 MHz and the −6 dB points are 14.03 MHz and 14.50 MHz as in fig. 3, then

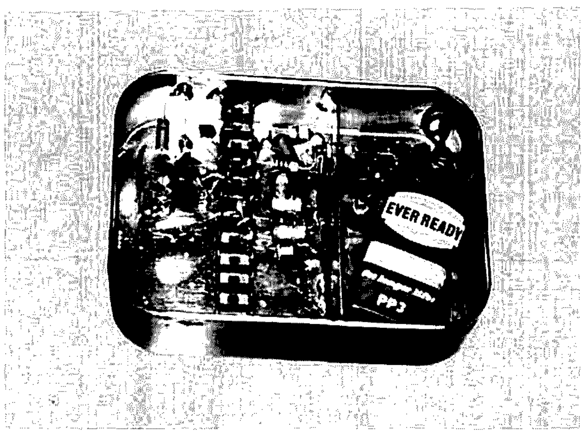
$$Q = \frac{170 \cdot 14.27}{100 \cdot 0.47} = 51.6$$

To measure loaded or in-circuit Q, it's possible to use the equipment's switching, coupling, and amplifying characteristics. For example, fig. 4 shows the input to a hypothetical multiband preselector with a coupled-pair of tuned circuits. The noise generator can be connected to the hot end of the first coil with capacitor C1 removed. To check the first coil, the noise generator can be connected to the antenna socket and a test receiver capacitively coupled to the hot end.

### response curves for simple filters

The shape of the frequency characteristic of the coupled pair in fig. 4 can be investigated with the noise generator connected to the antenna socket and the test receiver at the output (with C1 replaced). You then have the choice of a) plotting response in S-units using the receiver meter, or b) adjusting an attenuator inserted between the noise generator and the circuit under test to obtain a constant meter reading. This is an example of a simple filter easily checked with the noise generator, provided the receiver bandwidth is a few times narrower than the feature being investigated. A CW receiver will meet almost any needs; an ssb receiver will meet most.

The author's completed noise bridge. On the left is the actual noise generator circuitry.



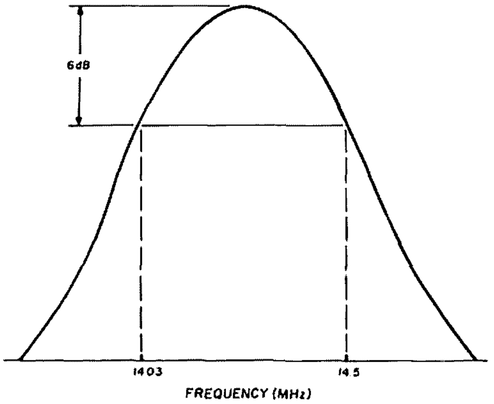



**fig. 3. Typical response curve obtained during in-place measurements to determine $Q$ of tuned circuits.**

### tuned-circuit alignment

The generator can be used to align any circuit whose wideband nature does not involve risks such as tuning an input circuit to the image frequency. Tuned circuits associated with crystal filters are particularly easy to align this way, as the need for accurate tuning of a stable generator is avoided — the crystal filter selects the frequency used.

### checking sensitivity

The sensitivity of a receiver may be checked by connecting the noise generator to its input through a switched attenuator. The generator is powered by ac and the attenuation increased until the buzz in the receiver output is only just distinguishable from the noise. This is a useful way of comparing receivers or checking circuit improvements.

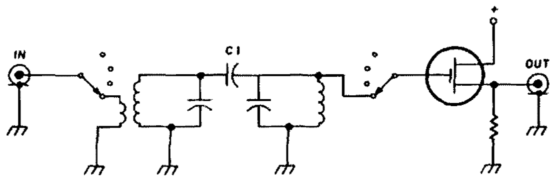



**fig. 4. An hypothetical preselector with a coupled pair of tuned circuits. The frequency response of the tuned circuits is easily checked with the noise generator of fig. 1 and a communications receiver of narrower bandwidth than that of the circuit under test.**

### design improvement

The Q values of homemade coils may surprise you — I found values around 25, and the generator has taught me to follow the rules:

1. Use the largest diameter coil and the largest diameter wire possible.

2. Space the turns by one-half to one wire diameter.

**ham radio**





More Details? CHECK—OFF Page 126

# 36-volt solar power source

## A system of jumpers and switches to allow versatile metering, maintenance, and use of solar-charged batteries

In my normal solid-state experimental activities and QRPP operations, several days pass before the 12-volt, 5.5 ampere-hour, sun-charged battery needs to be recharged (see the November, 1974, *Circuits and Techniques* column). Thus two additional motorcycle batteries have been included along with a new control panel, fig. 1. Power capacity has been increased and both 24 and 36 volts have been made available. These higher voltages are helpful for work with fets and higher-voltage bipolar transistors.

Each battery can be trickle-charged or charged at the 300-mA rate (in bright sun) each third day. Two or three batteries can be parallel-charged if desired, or other arrangements can be combined depending on your specific needs. Based on a 20-hour discharge rate the current available *from* each battery is

$$I = \frac{5.5}{20} = 275\ mA$$

Available power rates based on the 20-hour discharge time, using series-connnected units, are

1 battery = 12 x 0.275 = 3.3 watts
2 batteries = 24 x 0.275 = 6.6 watts
3 batteries = 36 x 0.275 = 9.9 watts

In parallel combinations of two and three batteries, power values are

2 batteries = 12 x (2 x 0.275) = 6.6 watts
3 batteries = 12 x (3 x 0.275) = 9.9 watts

The figures above are for conservative operation and are based on 20 hours of demand at 275 mA per battery.

How many hours per week do you operate QRP? Even more significantly, during those operating hours your maximum power demand would be intermittent, which will extend your operating-hour capability. The 12-volt, 300 mA solar panel can easily keep your batteries charged for these operating conditions. In my case, the solar power supply supports a considerable amount of solid-state experimentation in addition to on-the-air operation.

Let's look at power capability based on an 8-hour discharge rating

$$I = \frac{5.5}{8} \approx 700\ mA$$

Batteries in series

1 battery = 12 x 0.7 = 8.4 watts
2 batteries = 24 x 0.7 = 16.8 watts
3 batteries = 36 x 0.7 = 25.2 watts

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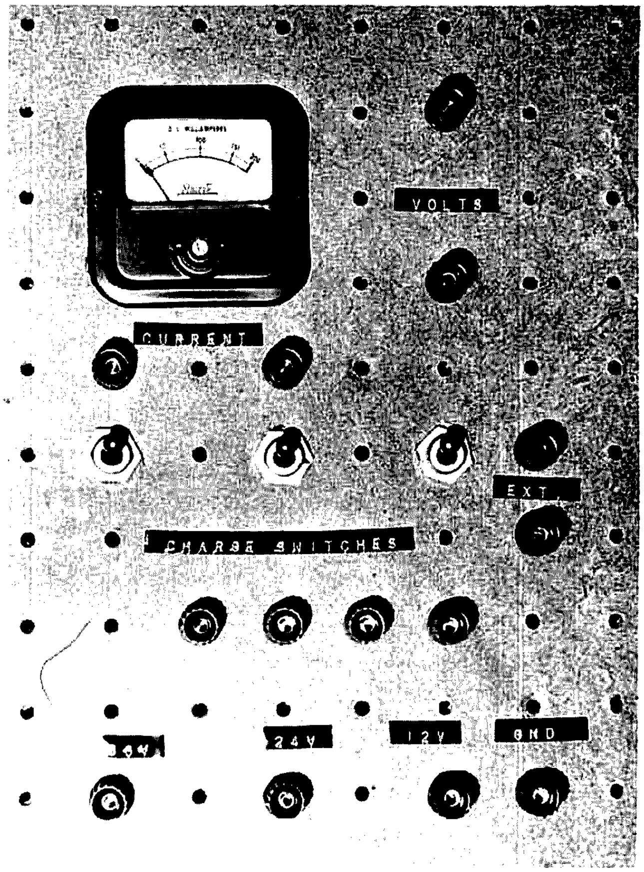



fig. 1. Solar control panel.

Batteries in parallel

2 batteries = 12 x (2 x 0.7) = 16.8 watts
3 batteries = 12 x (3 x 0.7) = 25.2 watts

The 8-hour figures indicate that you can draw considerable power from the 3-battery combination for extended periods of operation because of the very nature of the power demand during QRP activities. However, a much longer period is needed to restore the batteries to full charge, because the maximum charging rate is only 300 mA during maximum sunlight.

## working with the sun

Assume 36 hours of weekly QRP operation at 5 watts. Based on a normal operating situation, a conservative figure for CW and ssb operating time at maximum demand would be one-third this value or 12 (36/3) hours. What would be the weekly watt-hour demand? Watt-hours are a measure of the *quantity* of electricity and are the product of watts and hours

$$\text{watt-hours} = Wh = 5 \times 12 = 60\ Wh$$

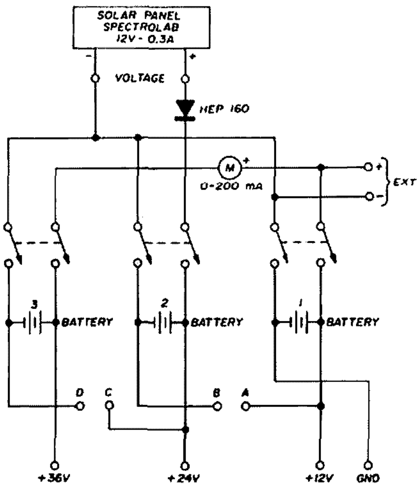
This amount of energy must be returned to the battery. In maximum sunlight the solar energy converter at W3FQJ supplies 3.6 watts (12 x 0.3). How many hours of a brightness level producing this amount of output would be required to replace the weekly watt-hour demand on the battery according to the preceding operating schedule? Assuming 100% efficiency

$$h = \frac{Wh}{W} = 16\ 2/3\ \text{hours}$$

This corresponds to approximately 2-1/3 hours of saturation-level brightness per day, averaged over a week's time. Indeed this amount allows very conservative operation of the solar power source. The one-third correction factor for normal operation is also conservative, compensating for the fact that the power demand on receive is substantially less than on transmit. Furthermore, even during transmit, no significant power demand is made during the spaces of code keying. For sideband transmission the peak demand is made only during the voice peaks; therefore the average power drawn is substantially lower. Consider also the fact that light energy can be converted to electrical watts at levels less than saturation brightness. Even during overcast and in the early morning and late afternoon hours, watt-hour replacement is being made. This extra charge time at lower levels more than compensates for the fact that you can't assume 100% efficiency in the battery-charging process. There is great emphasis today on the development of high-efficiency, lightweight and small-size batteries.

It is instructive to consider watt-hour capacity in terms of the solar panel itself. Again let's be conservative and assume the average brightness is such that saturation current is present something more than 3 hours per day. This average load figure more than compensates for days of dark overcast and rain. On this basis it is safe to assume approximately 25 hours per week of available saturation current, which corresponds to 90 watt-hours. If not maximum, at least significant power is being made available for 3 to 5 hours during the remainder of the day even during the winter months. Therefore the total weekly capability is at least 150 watt-hours. This is an average figure – some weeks more; other weeks less. The

fig. 2. Schematic of three-battery control panel.



actual figure also depends on geographical location. Consumption rates of this level, week after week, operate the system near limit at W3FQJ. This data indicates that 10 watts for 15 hours or 25 watts for 6 hours, averaged over the week, will not overburden the system. Using the one-third correction factor, this amounts to 30 operating hours at 10 watts or 18 operating hours at 25 watts.

The maximum power figures above are based on a weekly average. You must be careful not to demand too much power in any one day. For example, if you demand 25 watts for two hours in any one day (six hours of operation), your consumption would be 50 watt hours. *Recharge time at maximum sun would be nearly 14 hours (50/3.6).* Also in the same time you'll have used a bit more than 4 ampere-hours (25/12 x 2). This is a significant discharge of a 5.5 ampere-hour battery. *Another hour of such continuous operation without recharge would discharge your battery completely.* Continuous and average demand are factors in solar power systems. Each must be considered individually when working a system near its limit.

## new control panel

The new control panel has been designed for maximum operating convenience here. You may develop plans that better meet your specific needs. However, the description will give you an idea of how the panel can be arranged for versatile charging and use. Three charge switches permit the batteries to be charged one at a time or two or three simultaneously. This is an aid in dividing the charge current as a function of individual battery use. A 0-200 dc milliammeter is connected in the charge path to the first battery only. It reads the current delivered from the solar panel when battery 1 only is under charge. If the charge switches are closed to the other batteries, this reading will decrease and will give some idea of how current divides among the three batteries. Remember, too, that when parallel-connected, batteries charge each other. Furthermore, a bad battery in the combination can place a heavy load on a good battery.

Batteries can be connected in series or parallel depending on the operating voltages to be established. Another pair of binding posts is used for external battery charging directly from the solar panel. An additional pair of binding posts permits a measurement of the solar-panel voltage at the point where the incoming wires connect to the control panel. The schematic is shown in fig. 2. Note that the protective-diode cathode and common connect to the poles of the dpst switches. The batteries, using associated binding posts, are connected across the individual terminals of the three switches. Therefore 1, 2, or 3 batteries can be charged at the same time. Battery 1 has its negative side connected to ground. 12 volts can be obtained by connecting the load between +12 volts and ground. Whenever 24 or 36 volts is used all three charging switches are turned off. To obtain 24 volts a jumper is connected between binding posts A and B, which makes +24 volts available at the output binding post labeled +24V. When output is to be obtained from the +36-volt binding post, jumpers must be connected between binding posts A and B and between C and D. *Never operate the charging circuit with any jumpers connected between A and B or C or D.* A battery short circuit will result if either the battery 2 or battery 3 charge switches are closed with either of these two jumpers in position.

Parallel operation of the three batteries can be obtained by joining the +12, +24, and +36 volt binding posts together. Additionally a ground is established by

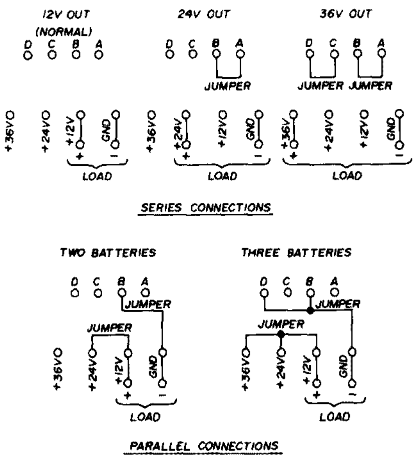



**fig. 3. Use of jumpers in making series and parallel battery connections.**

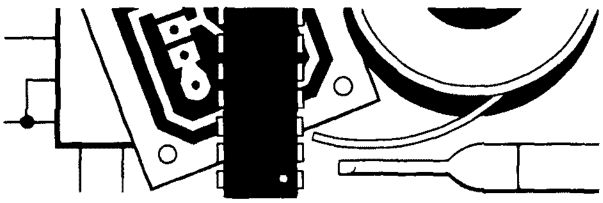
joining binding posts D and B with ground. The various binding post interconnections for various series and parallel groupings of the batteries is detailed in **fig. 3.** The 12-volt single-battery operation is inherent. The additional batteries in series or parallel require the use of appropriate jumpers. The charging of an external battery requires only that it be connected across the EXT binding posts. With all charge switches off, the meter will read charging current to the external battery.

## expro I on forty meters

The Expro I fet transmitter in the August, 1975, column was adapted for 40-meter operation by winding a separate toroid coil. No other charges were required. The coil was wound on an Amidon T68-2 core and consists of 40 turns, number-24 enameled copper wire close-wound. It tuned up well on the antenna also described in the August column. The first contact I made on 40 meters was a special thrill. It was made with "Pop," W2ZTC, of Hancock, New York. Pop is an 88-year old gent with a hand-key CW rhythm better than my own. Thank you Pop for your patience with my rst 539 weak signal. You're the most! Your homebrew 100-watt rig sounded great. This is an example of amateur radio joining the generations.

**ham radio**

## the weekender



# a simple computing vswr indicator

The reasons for using a transmatch to match a random-length wire antenna to a typical 50-ohm transmitter output impedance are well known. However, practical experience shows that adjusting the transmatch for a good match can become a tedious affair. Since transmitter output changes with the load presented to it, you must constantly put the typical swr bridge on *forward,* adjust for a full-scale meter reading, then switch to *reflected* to read the swr. Otherwise, if you leave the bridge on *reflected* and tune for minimum meter indication, you usually find that, instead of being tuned for minimum swr, you've tuned for minimum transmitter output power instead.

The swr meter described here makes transmatch tuning incredibly simple, since it automatically uses the ratio of forward and reflected wave components to display the swr continuously regardless of transmitter output power. All you have to do is tune the transmatch to minimum meter deflection, thus avoiding the constant switch flicking and pot diddling usually required with swr meters. The circuit has some advantages over other types of computing swr meters[1,2] including:

1. A minimum of parts, which means a minimum of time and money to put it together.
2. Conventional power requirements. The unit may be powered from a 12-volt battery or from a ±9 or ±12-volt supply.
3. Accuracy over a 1000:1 range of transmitter power levels with no sensitivity adjustment needed. Actually a 100:1 or even a 20:1 range would be adequate for most situations.
4. Adaptation to existing swr bridges.
5. Only two adjustments needed for initial calibration.

### operation

To understand how the circuit works, we should first think about how the typical swr meter operates. The usual swr indicator has a directional coupler inserted into the coax line. (The coupler could be of the Varimatch variety,[3] or it could be a directional wattmeter[4]). The coupler outputs a pair of dc voltages proportional to the forward and reflected voltages in the coax. When you adjust the swr meter sensitivity so that the forward dc voltage produces full-scale deflection, you have essentially calibrated the meter; applying the reflected dc signal to the meter allows you to read the swr directly. What this computing swr meter circuit does is

*A complete parts kit for this computing vswr indicator is being made available in conjunction with this article. For ordering information and prices, write to G.R. Whitehouse & Co., 10 Newbury Drive, Amherst, New Hampshire 03031.

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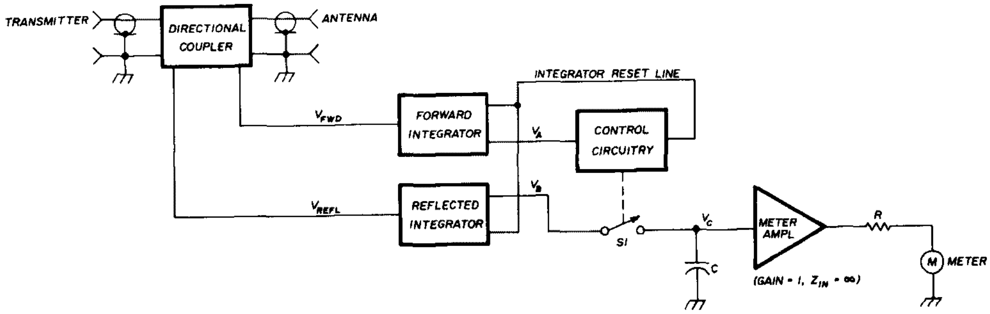
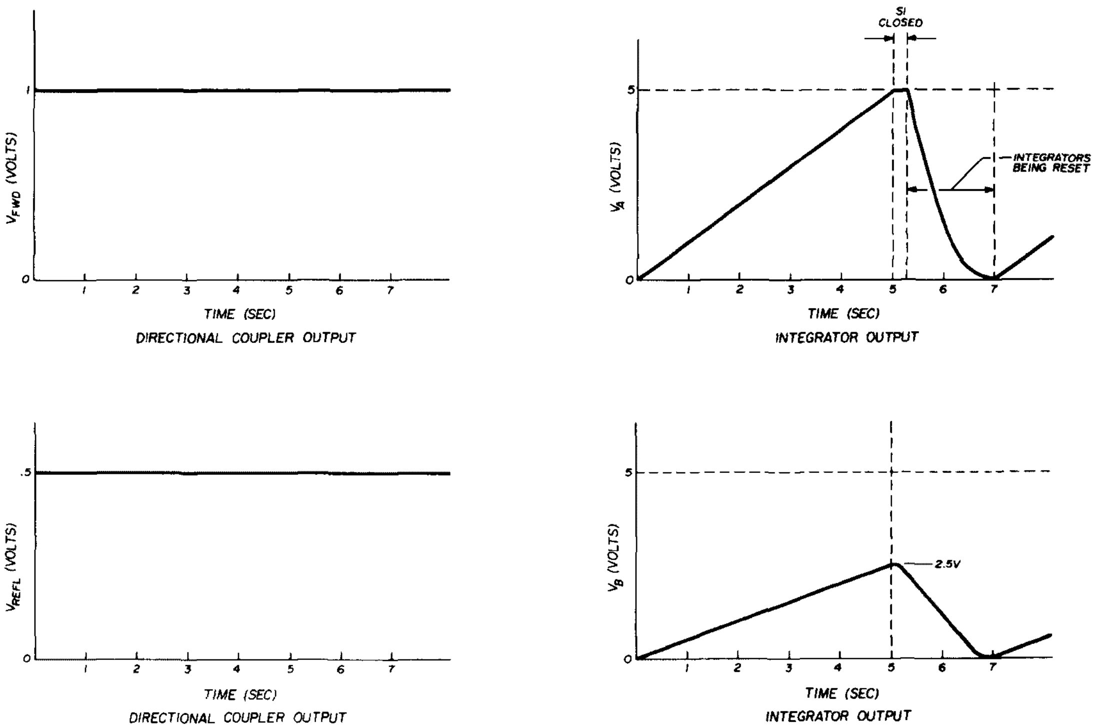
adjust the meter sensitivity automatically as it applies the reflected signal to the meter.

Referring to the block diagram of **fig. 1** and its timing diagram, you can see that the directional coupler output voltages are applied to a pair of integrators. The integrator outputs are ramp waveforms for constant inputs; the higher the input voltages the steeper the ramps become. At any instant in time, the ratio of integrator outputs is



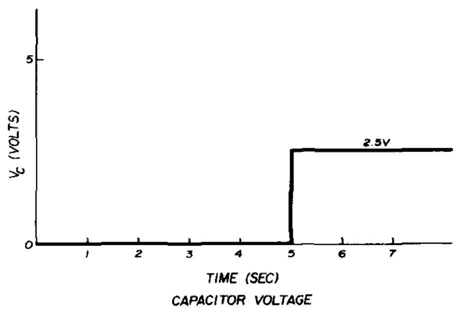




**fig. 1. Block diagram of the computing swr indicator, above, and its timing relationships, below. A pair of integrators form the heart of the system. At any instant of time the ratio of their outputs is equal to that of the input voltage, assuming the input-voltage ratio is constant, i.e., constant vswr.**

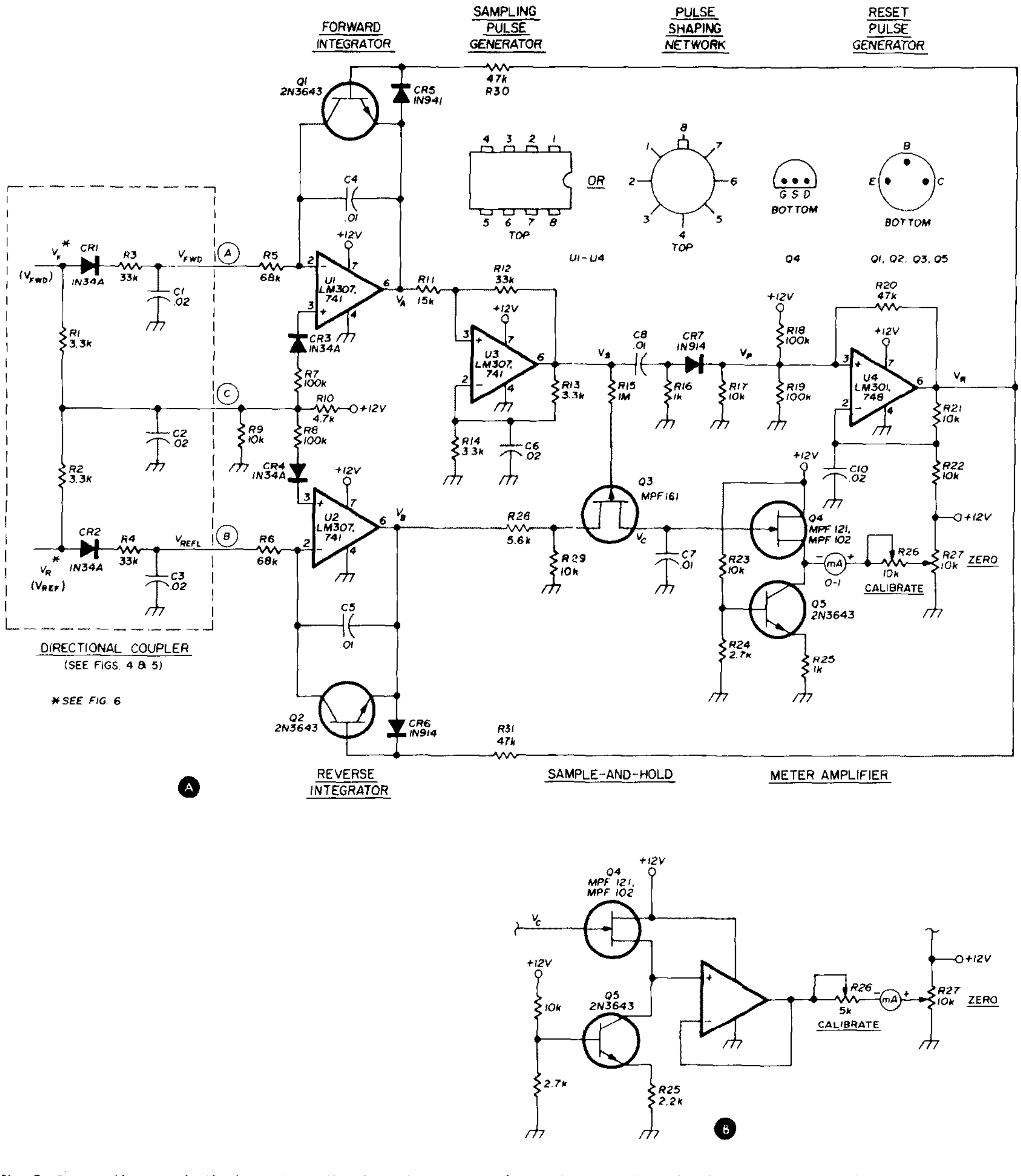



**fig. 2. Computing swr indicator schematic, A, and an alternative meter-amplifier circuit, B. If a single 12-volt power supply is used, a special directional coupler is required (see fig. 4).**

equal to the ratio of their input voltages (assuming the *ratio* of the input voltages remains constant; i.e., constant swr).

Now, let's say that *R* in **fig. 1A** is set so that 5 volts at $V_C$ produces full-scale meter deflection. The integrators produce ramps $V_A$ and $V_B$. When $V_A$ reaches 5 volts the control circuit briefly closes S1 so that the capacitor is charged to $V_B$. $V_B$ is 2.5 volts at this time. The meter is now at half-scale deflection, indicating an swr of 3:1. After opening S1, the control circuit resets the integrator outputs to zero and the process starts all over again.

The capacitor will hold the 2.5 volts until S1 is again closed, because the meter-amplifier input draws essentially no current.

In the circuit of **fig. 2A** (and the corresponding timing diagram of **fig. 3**), you can see that the integrators operate around +8 volts, instead of zero volts, when you use the single +12 volt supply shown; this requires that a special directional coupler circuit be used (see **figs. 4** and **5**). If a conventional directional coupler is desired, you must use a dual-polarity power supply as discussed later. When $V_A$ drops to 3.5 volts, the positive input of

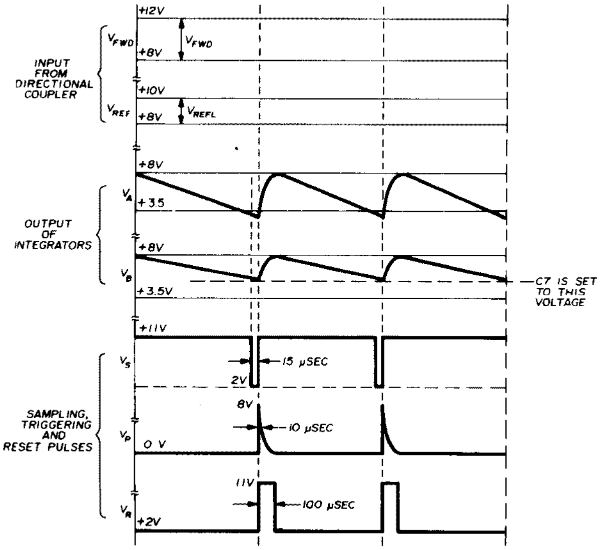



**fig. 3. Timing diagram for the circuit of fig. 2A showing relationships between input signals, integrator outputs, and sampling, trigger, and reset pulses.**

U3 is pulled below the voltage at the negative input, forcing the output low. It is held low until C6 discharges to where the negative input is closer to ground than the positive input, forcing the output high. When the output is low, Q3 is turned on, charging C7 to a new voltage.

The rising edge of U3 output is applied to U4, which generates a pulse in a similar manner, resetting the integrators to 8 volts by discharging C4 and C5 through Q1 and Q2. CR5 and CR6 prevent the transistor base-emitter junctions from accidentally breaking down and turning on the transistors when U4 output is low.

The meter amplifier, Q4 and Q5, can drive meters with sensitivities of 1 mA or better. If the meter you happen to have has lower sensitivity than this (or if you want *good* meter accuracy) then you should probably use the meter-amplifier circuit shown in **fig. 2B**. In the circuit of **fig. 2A**, Q5 acts as a constant 2-mA current source; more meter current means less current through Q4, which would mean a change in Q4 gate-source voltage, introducing some dc error. In **fig. 2B**, the change in Q4 current is negligible.

The meter is calibrated using the usual swr meter scale. You can calibrate a junk-box meter by:

$$swr = \frac{I_{fs} + 1}{I_{fs} - 1} \qquad (1)$$

where $I_{fs}$ is the meter full-scale deflection current.

## construction hints

U1 and U2 should be good-quality op-amps requiring very small input bias currents. LM307s are a good compromise between cost and results. Type 741s could be used if you have some on hand, although they are slightly inferior to the less-expensive LM307. U3 and U4 should be uncompensated, since they are used as fast pulse generators. Internal compensation makes op amps such as the LM307 and the 741 too slow. Don't use 709s, because their differential input voltage ratings will be exceeded in this application.

Diodes CR3 and CR4 provide better sensitivity at low transmitter power levels by compensating for the forward drop across CR1 and CR2. Diode pairs CR1-CR3 and CR2-CR4 should be matched for forward-voltage drop at low currents. (Do this by matching their forward resistances on the x10k and x100k scales of a vtvm). Diodes CR1 and CR2 should have low reverse leakage current (high back resistance). Using silicon instead of germanium diodes decreases circuit accuracy at low rf power levels (10 watts and less).

Q1 and Q2 should have low leakage current through the collector-base junction; most silicon transistors will work fine here. Check the $I_{dss}$ of Q4 if you wish; if it's 3 mA or more, decreasing R25 to 680 ohms will reduce the meter-amplifier error mentioned earlier, since Q4 will then supply 3 mA. Don't worry about this if your meter sensitivity is better than 500 μA; however, be sure to increase R26 to a value that will protect the meter movement.

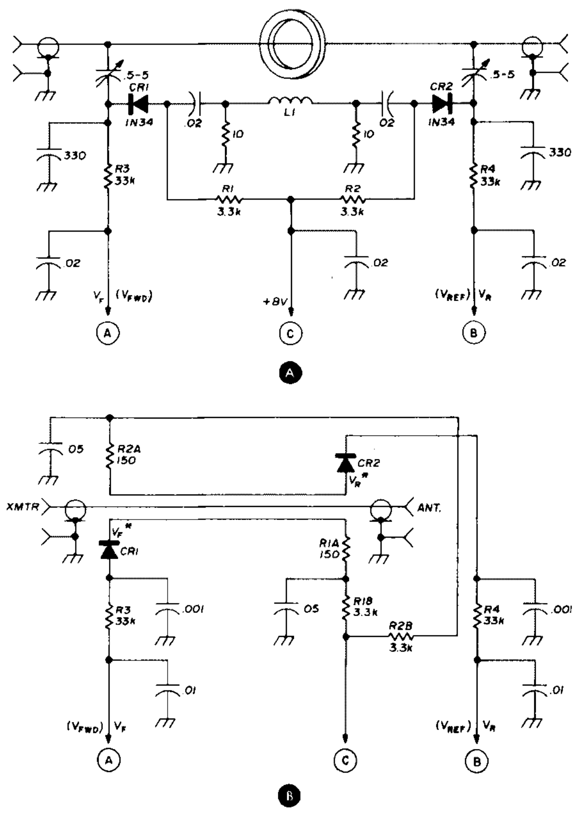



**fig. 4. Circuits to be used in modifying existing swr bridges for use with the computing swr meter. A modified directional wattmeter is shown in A; a modified Monimatch in B. Inductor L1 consists of 30 turns of no. 22 AWG (0.6mm) wire on an Amidon T-50-2 form.**

If you build this circuit into an existing swr meter, you can use **figs. 4** and **5** as guides when wiring the directional coupler to the circuit of **fig. 2A**. In general, expect better results if you use a Varimatch coupler rather than a directional wattmeter. Directional wattmeters don't seem to work well at power levels below 5 or 10 watts because of their low sensitivity. If you build your circuit from scratch, the directional coupler shown in **fig. 5** is easy to build and gives good results to at least 30 MHz.

If you have a dual-polarity power supply (±9 or ±12 volts), you'll probably want to use a conventional directional coupler circuit. If so, connect point C of **fig. 2A** to ground and connect all other ground to the negative supply line. Also, when using either single-polarity supplies of more than 15 volts, or dual-polarity supplies, you'll probably want to increase R23 and R26 (and possibly R27 as well). Except for these minor changes, the circuit will work as is on any voltage between 12-28 volts. At 12 volts the circuit draws approximately 20 mA, making a battery power supply practical. Since the meter calibration accuracy depends on the supply voltage, extreme accuracy will require the use of a regulated power supply. However, as long as the power-supply voltage doesn't vary by much more than ±10% or ±15%, the resulting accuracy will be more than adequate for practical use. Be sure that no strong rf fields can get into the op amps. Install the circuit in an aluminum box. Filter all leads entering and leaving the box.

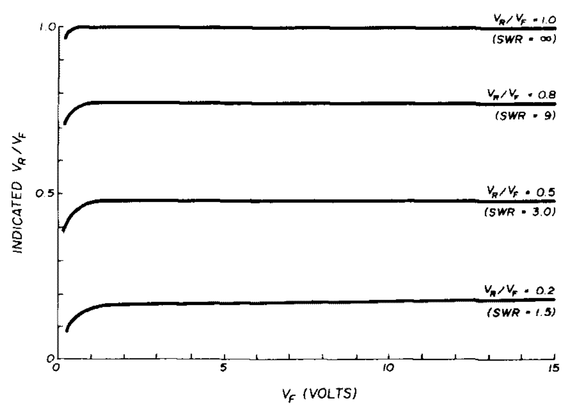
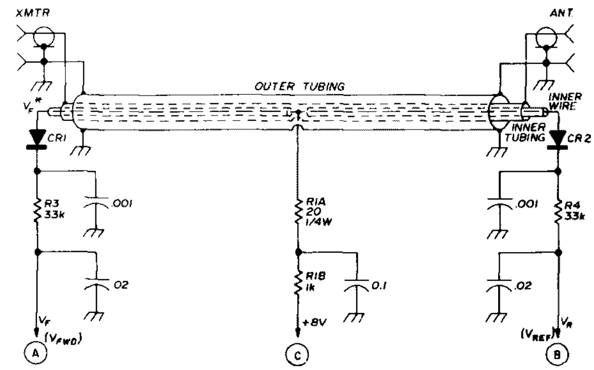



**fig. 6. Actual vs indicated ratios of $V_R/V_F$ as a function of $V_F$. $V_F$ and $V_R$ were dc voltages applied to points $V_F$* and $V_R$* of fig. 2A. CR1-CR4 were 1N914s; R5-R8 were 1 megohm. By using the components specified in fig. 2A the curves should extend in nearly straight lines to $V_F$ = 0.1 volt or so.**

## calibration and results

The first step in calibration is to check that the directional coupler is working properly. Connect a dummy load to the antenna jack and check for zero volts between points $V_{refl}$ and C of **fig. 4** or 5. After removing the dummy load, you should get nearly equal voltages at $V_{fwd}$ and $V_{refl}$ (they don't have to be exactly equal). These steps need not be taken if you use a dual-polarity power supply and an old but reliable swr bridge directional coupler.




| | |
|---|---|
| **Inner conductor** | inner part of RG-59/U coax with insulation, prepared as shown above |
| **Inner tubing** | 1/4" (6.5mm) diameter copper tubing, 8" (20cm) long, with a small hole (1/8" or 3mm) filed at the midway point |
| **Outer tubing** | 1/2" (13mm) copper tubing, 7" (18cm) long, with a small hole filed at the midway point |

**fig. 5. The Varimatch circuit modified for use with the computing swr meter of fig. 2A.**

Set the *calibration* pot (R26) for maximum resistance and apply power to the circuit. With the dummy load connected and the transmitter key down, adjust R27 for zero meter deflection. Next, disconnect the load from the antenna jack, briefly close the key and adjust R26 for full-scale meter deflection. You may want to touch up R27, but this probably won't be necessary. Your computing swr meter is now calibrated and ready for use.

If the circuit doesn't work properly, first check that the *transmitter jack really is* the transmitter jack. Then, by using very low transmitter power (a half watt or less),

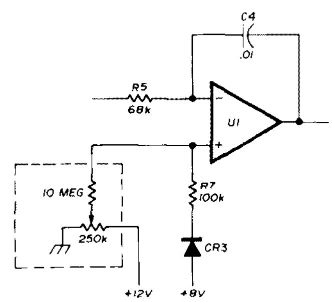



**fig. 7. Optional circuit for adjusting U1 offset voltage if the swr meter indicates a random value during key-up conditions. So long as U1 output remains above 3.5 volts the meter needle will stay put with no transmitter output.**

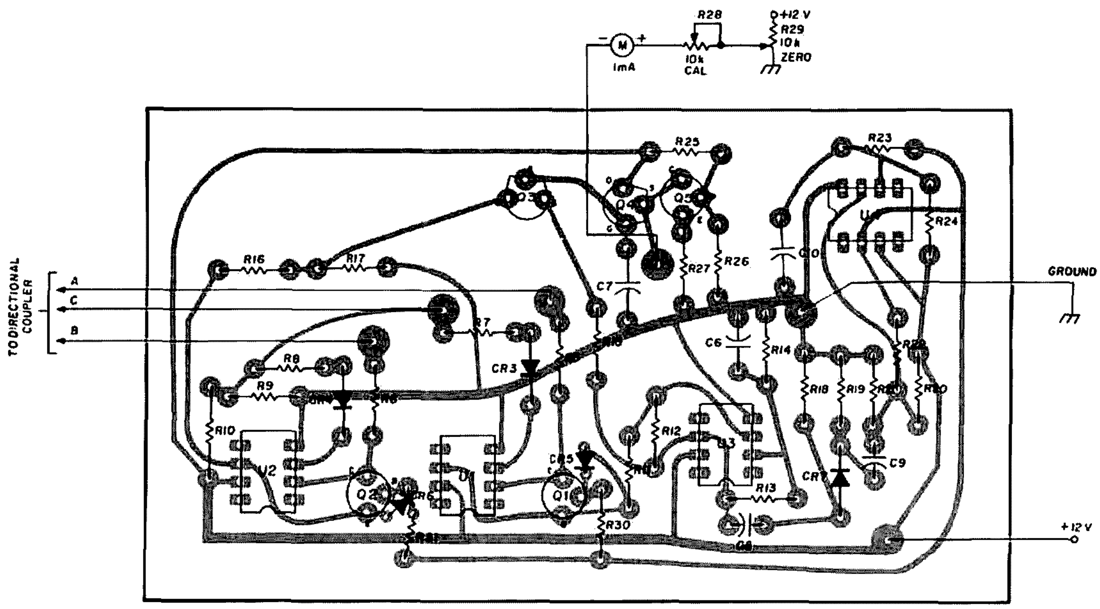



**fig. 8. Printed-circuit component layout for the swr computer. A printed-circuit board for this project is available from G.R. Whitehouse (see footnote on page 58).**

you should be able to see the integrators move with your vtvm. Pulses at U3 and U4 can be detected with the vtvm set to ac volts or with a vom set to *output.*

Typical performance using silicon diodes for CR1-CR4 is shown in fig. 6. The integrators are almost impossible to overload (an overload would correspond to trying to make the outputs move faster than 0.5 volt per microsecond), so the upper limit to what power the circuit will handle is set by the PIV of CR1 and CR2 and by the width of the pulse fed to Q3. The lower limit for good meter accuracy is set by how well diodes CR1-CR4 are matched, plus whether the diodes are germanium or silicon. The constant offset between indicated and actual $V_R/V_F$ ratio is due to error in the meter amplifier.

Fayman[1] noted that his swr meter had a tendency to sit at random values of swr during key-up conditions, requiring special circuitry to correct the problem. The circuit of fig. 2A can have the same problem, but it can be overcome by adjusting U1 offset voltage (see dashed box, fig. 7) so that its output drifts slowly toward the positive end of the supply line with no transmitter output. (Check your circuit before adding the pot and resistor of fig. 7; your particular U1 may already be doing this). As long as U1 output remains above 3.5 volts, Q3 will remain turned off, and the meter needle will stand still (it won't go to zero when the key is up, but it won't go off to some random value either).

## concluding remarks

This computing swr meter has been a big help in transmatch tune up. It also has suddenly made all kinds of antenna ideas appear appealing since the transmatch is now an easily used piece of equipment. Even so, many improvements could be made in this circuit. You could, for instance, try a circuit allowing the use of *both* a conventional directional coupler *and* a single-polarity 12-volt supply. This could be done by using CA3130 CMOS op amps for U1 and U2. Diodes CR1 and CR2 would have to be turned around, and CR3, CR4, R7 and R8 would have to be eliminated; also, an amplifier would have to be used between U1 and U3 to transform a signal exceeding +10 volts to a signal dropping below +3.5 volts.

You could also try reducing the cost (at a sacrifice of accuracy) by using an LM3900 quad amplifier. The problem here is that you'd have an swr of 1:1 every time $V_{refl}$ dropped below 0.7 volt or so – not so neat if $V_{fwd}$ is only 0.8 volt zero to peak. However, for higher voltage levels, this idea might be OK. Another idea would be to add a peak-reading circuit to the directional coupler $V_{fwd}$ line to indicate transmitter PEP in sideband applications.

## references


1. D. Fayman, W0GI, "A Simple Computing SWR Meter," *QST,* July, 1973, page 23.
2. T. Mayhugh, W6OTG, "A Digital SWR Computer," *73,* November, 1974, page 80; December, 1974, page 86.
3. D. DeMaw, W1CER, "The Varimatcher," *QST,* May, 1966, page 11.
4. D. DeMaw, W1CER, "In-line RF Power Metering," *QST,* December, 1969, page 11.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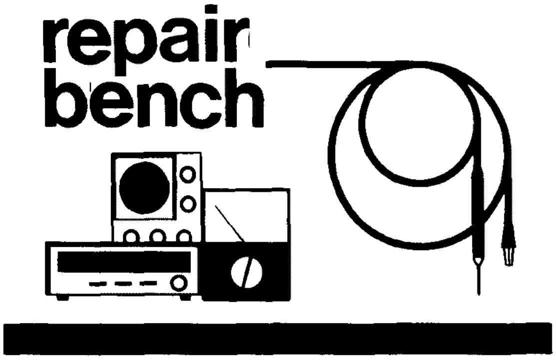


# Bob Stein, W6NBI

## using the swr indicator

Ever since the end of World War II, the surplus market has been a major source of test equipment, transmitters and receivers, and parts. For the serious experimenter, the first category has been of paramount importance, enabling him to obtain equipment which would otherwise be out of the question because of cost. Surplus equipment continues to be in strong supply from both industrial and military sources. In fact, there is more available now than ever before, for several reasons.

First of all, the state of the electronics art has changed so rapidly in the past few years that much of the test equipment which was standard in industry is no longer economical to use, although it is perfectly adequate for amateur use. A typical example of this is the slotted line used with the swr indicator, a combination that has largely been replaced by the reflection bridge or the network analyzer. A second reason for the availability of good commercial test equipment is the fact that the military services have been using off-the-shelf commercial equipment, which they continue to surplus.

This article is the first of a series which will explain the uses and benefits of certain classes of test equipment which heretofore have been relatively uncommon in the ham shack. I hope to show how modest investments in such equipment will enable you to improve your testing capabilities as well as expand your knowledge of test methods.

### the swr indicator

Although there are and have been several types of swr indicators available, the one which is most common and which seems to be the easiest to obtain is the Hewlett-Packard model 415B Standing Wave Indicator. Others which appear on the surplus market from time to time are the Sperry Microwave SWR Indicator model 29A1 and the Narda model 441 VSWR Amplifier. Current instruments of the same type are the Hewlett-Packard model 415E SWR Meter and the GenRad (formerly General Radio) Standing-Wave Meter type 1234. Obviously there are differences among all of these, but basically they are similar.

Despite the similarity in names, the swr *indicator* to be discussed in this article is something completely different from the vswr *bridge* commonly found in most ham shacks. The bridge which you use in your transmission line is an rf *power* instrument which measures incident and reflected power or voltage (with varying degrees of accuracy). The vswr is then calculated from these readings, or is read directly from the meter provided that a reference level for the incident wave has been established.

The swr indicator, on the other hand, is an instrument which is exactly what its name says it is – an indicator. It is capable of indicating vswr only when used in conjunction with an rf detector and other equipment, such as a slotted line or a directional coupler. However, it can do this with no more power applied to the device under test than is produced by a signal generator. Thus, vswr readings can be obtained for virtually any type of equipment – converters, attenuators, filters, transmitters, coax relays and switches, antennas, transmission-line sections, and so on.

But that's only the beginning. The swr indicator can also be used with a detector to measure the attenuation and insertion loss of fixed and variable attenuators, insertion loss and response of filters, cross-talk in coaxial relays and switches, and antenna gain (when compared to a standard gain antenna on an antenna range). It can also serve as an extremely sensitive null indicator in ac bridge circuits. So you can see that we are talking about a very versatile instrument.

### inside the swr indicator

Despite its versatility, the swr indicator is a relatively simple instrument; it is basically nothing more than a tuned electronic voltmeter which responds to a fixed frequency of 1000 Hz. (There are special cases where a frequency other than 1000 Hz is used, but these are rare, and the instrument model number generally will have been modified by the addition of an option or special number. Check the back of the instrument housing for a sticker which shows this modified model number.) The gain is extremely high, resulting in a sensitivity of typically 0.1 microvolt for full-scale meter deflection on the most sensitive range.

The swr indicator incorporates range attenuators, a gain control, and provisions for selecting and matching the type of detector used. The meter scales are marked in swr and dB. Note that the presence of a gain control indicates that the instrument is not a calibrated voltmeter, but is one which is designed to measure the ratio of two voltages. This capability is all that is required, since swr and dB are, by definition, both ratios.

### outside the swr indicator

As mentioned earlier in this article, a detector must be used with the swr indicator for all rf measurements,

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and obviously an rf signal source at the desired frequency is required. The signal source must be amplitude modulated with a 1000-Hz sine or square wave. An output of 0.2 *milliwatt (100 millivolts* across 50 ohms) is generally sufficient. Regardless of the actual output power, the signal source must supply a constant output and should have minimum harmonic distortion. The modulating voltage must also be stable to prevent measurement errors.

The detector must be a square-law device, that is, one whose *output voltage* is proportional to rf *input power*. Coaxial slotted lines usually have a detector built into them, but an external detector must be used for other applications of the swr indicator. Such a detector may be either a coaxially mounted bolometer* or semiconductor diode, operated at low signal levels.

Although the Hewlett-Packard 415B can supply either 4.3 or 8.7 milliamperes bolometer bias current, bolometer detectors are relatively rare so we will consider all operation using a diode detector. There are many types, made by many manufacturers, all of which will be more than satisfactory for use with the swr indicator. However, be sure that the detector which you use presents a termination which corresponds to the system impedance; both 50- and 75-ohm detectors are available. For the purpose of this article, all impedances are assumed to be 50 ohms.

**The Hewlett-Packard model 415B Standing Wave Indicator. Although this instrument has been superseded by the model 415E, *it will still provide years of useful service.* (Photo courtesy of Hewlett-Packard Company)**


A coaxial detector is also a simple and inexpensive device to construct; methods of construction are shown in previous articles by W6VSV[2] and W1JAA.[3] Virtually all commercial detectors incorporate a resistive termination at the rf input connector, but those described in W6VSV's article do not. Therefore a 50-ohm feedthrough termination should be used at the input of such a detector. (Alternatively, a 50-ohm pad and coaxial short can be used, as shown by W6VSV.)

Some diode detectors incorporate a matched load resistor either at the *output* connector or in a separate coaxial load. These too are relatively rare, but bear mentioning. Their advantage is greater square-law range, although this is achieved at a cost of some sensitivity loss.

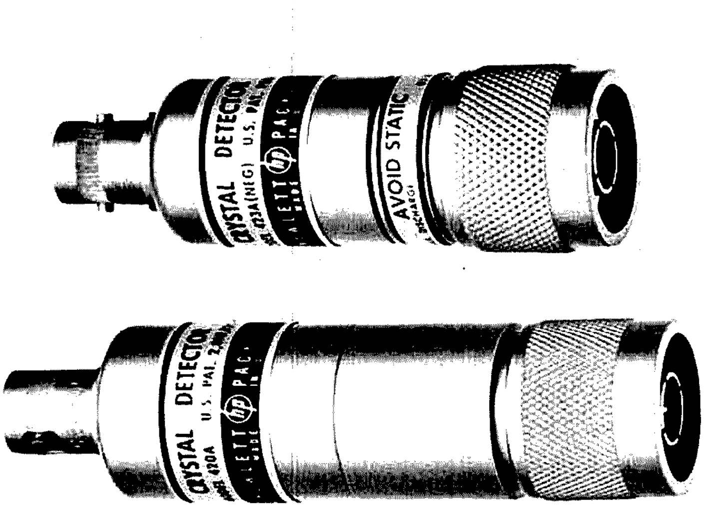
There are two precautions which must be observed when using a diode detector. The first of these entails staying within the square-law region of the diode. For amateur measurements, power levels up to a few microwatts may be applied to the detector input without causing serious measurement errors. In general, if the applied rf power is kept below the level which results in a full-scale reading on the 30-dB range of the swr indicator with the gain controls at maximum, the detector will be working within its square-law range.

Secondly, if a detector with a matched load is used, the input selector switch on the swr indicator must be set to the high-impedance (200 kilohm) *crystal* position in order not to shunt the detector load with a low impedance. As previously stated, however, detectors with matched loads are relatively uncommon. With an unloaded detector, the input selector is set to the *crystal* position which results in the highest sensitivity.

The attenuation of loss pads, cables, filters, etc., may be measured using the test set-up shown in fig. 1. Loss pads, preferably 10 dB or more, are used at the input and output of the device under test (DUT) to ensure a uniform 50-ohm system. If the detector presents a known 50-ohm impedance at the test frequency (as most commercial units do), the pad between the DUT and the detector may be omitted.

Connect the equipment as shown, except substitute a feedthrough coaxial fitting for the DUT. Adjust the signal-generator output and the range switch and gain controls of the swr indicator for a reference level of 0 dB on the lowest range (most counter clockwise range-switch setting) which will allow for a reduction of signal level equal to the expected attenuation of the DUT. This ensures operation of the detector in its square-law region. For example, if a 20-dB attenuator is to be checked, setting the swr indicator reference level to 0 dB on the 40-dB range setting will allow you to measure up to 30 dB of loss – two 10-dB ranges of the switch plus the 10-dB meter range.

*There are two types of bolometers: *barretters*, which are normal resistance elements with a positive temperature coefficient, and *thermistors*, which are manufactured from metallic oxide materials which exhibit a negative temperature coefficient. In its simplest form the barretter may be a short length of very fine wire, such as an instrument fuse, or a metallized film resistor.[1]



**The Hewlett-Packard models 420A and 423A Crystal Detectors are typical of the many detectors which can be used with the swr indicator. Both detectors are usable from 10 MHz to 12.4 GHz. The response of the 420A, on the top, is ±3.5 dB with a maximum vswr of 3.0. The 423A is flat within ±0.5 dB and has a maximum vswr of 1.5 (1.2 below 4.5 GHz). (Photo courtesy of Hewlett-Packard Company)**

Insert the DUT into the test set-up, in place of the coax feedthrough, and note the swr indicator meter reading. *Do not touch the gain controls.* If the meter is still on scale, the attenuation of the DUT is equal to the meter reading on the dB scale. If the meter reads below the 10-dB mark, switch to the next lower (counterclockwise) range-switch setting which moves the meter on-scale. The attenuation of the DUT in dB then will be equal to the difference between the reference and final switch settings plus the meter reading.

Continuing with the example cited above, assume that when the 20-dB pad is inserted in the test set-up, the range switch has to be changed from the 40- to the 60-dB position, and the meter pointer rests at the 1-dB mark. The attenuation of the loss pad is thus actually 21 dB – two 10-dB switch positions plus 1 dB on the meter.

The Hewlett-Packard 415B has a *meter scale* switch which makes reading the meter more accurate under certain conditions. Since the dB scale is logarithmic, the meter scale between 5 and 10 dB is greatly compressed. If the meter reading is in this region, you can set the *meter scale* switch to the *−5 dB* position and turn the range switch to its next lower (counterclockwise) setting. The meter pointer will then fall between 0 and 5 dB, providing a more accurately read indication, but don't forget to subtract 5 dB from your final readings.

If a reading of less than 2 dB is involved, such as would hopefully be expected when measuring insertion loss, it can be read more accurately on the *expanded dB* scale of the meter. This requires setting the *meter scale* switch to *expand* and establishing the reference level with the *meter scale* switch in that position.

## measuring filter response

The attenuation vs frequency response of a filter may be measured in a manner similar to that just described. However, certain additional precautions are involved. First and foremost, your signal source output must be kept constant over the frequency range of interest. Secondly, if you are interested in the response below −35 dB, the loss pad at the input to the filter must be adjustable or be changed, since 35 dB is about the limit of a typical diode detector's square-law range.

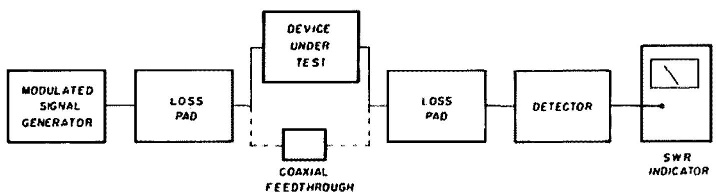
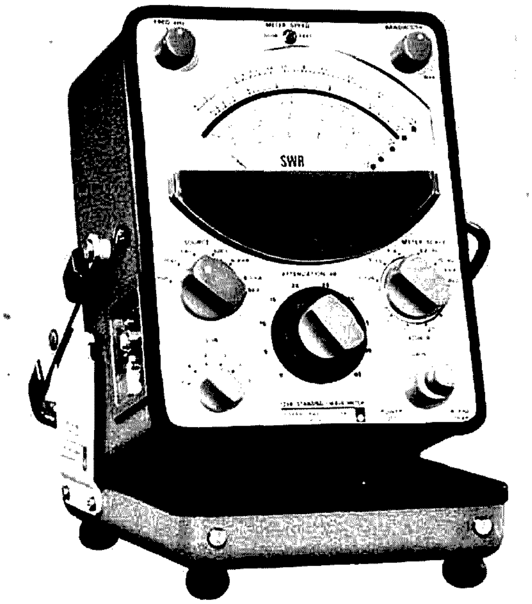



**fig. 1. Equipment set-up for measuring attenuation, filter response, and gain. The coaxial feedthrough connector is used to establish a reference level on the swr indicator, and is replaced by the device under test (DUT) for the actual measurement. The signal generator may be replaced by any stable small-signal source which is modulated by 1000 Hz signal.**




The type 1234 Standing-Wave Meter is the current swr indicator in the GenRad (formerly General Radio) product line. (Photo courtesy GenRad)

As an example, let's assume that you want to check the response of a 145-MHz bandpass filter. Connect it as shown in fig. 1, using a frequency counter at the output of the signal generator if the frequency calibration of the generator is in doubt. Instead of a fixed loss pad between the signal generator and the filter, use a step attenuator of known accuracy, set to provide 40 dB of attentuation. Tune the signal generator for maximum indication on the swr indicator, using the 30-dB range with the gain controls at or near maximum. Establish a 0-dB reference level by means of the signal generator output attenuator and the swr indicator fine-gain control.

As the generator frequency is varied, the attenuation of the filter can be read directly from the dB scale of the meter in conjunction with the range switch. As you get further away from the center (reference) frequency, you will run out of swr indicator sensitivity. At this point, note the meter reading, reduce the sensitivity of the swr indicator by 20 dB, and lower the attenuation of the step attenuator by 20 dB, thus still keeping a 20-dB pad in the system for impedance matching. If your step attenuator is accurate, the meter indication should be the same as before; if it is not, it indicates an error in the step-attenuator calibration which must be accounted for in subsequent readings. Now continue to check the filter response, adding 20 dB (or the amount by which the step attenuator was changed) to the swr meter readings.

If you don't have a step attenuator, you can accomplish the same thing by using two fixed loss pads between the signal generator and the filter. When you reach the point of maximum swr indicator sensitivity, simply remove one of the pads and continue as described above. In this case, you must add to the swr indicator readings the attenuation of the pad which was removed.

An immediate reaction to these procedures may well be that the accuracy of the measurements at low levels depends on the calibration of the adjustable or additional fixed attenuator. This is true, but all you have to do is to *measure* the attenuator, using the same swr indicator and detector, if you are not sure of the calibration. What could be simpler than a self-checking system?

## measuring gain

It should be obvious that if attenuation, or signal loss, can be measured with an swr indicator, signal gain can also be determined. This is accomplished by exactly the same procedure described for attenuation measurement, except that the reference level through the coax feedthrough should be established on the most sensitive (60-dB) range of the swr indicator.

For example, let's assume that you want to know the gain of your two-meter converter, and expect it to be between 25 and 30 dB. If you set the 0-dB reference level on the 60-dB range of the swr indicator, you will have to reduce the sensitivity when the converter is connected in place of the feedthrough connector. The amount by which sensitivity is reduced is equal to the change in the range-switch setting *minus* the meter reading, and represents the gain of the converter.

Conversely, gain can be measured by establishing a reference level using the device under test. Then substituting a coaxial feedthrough for the DUT will result in a loss equal to the gain of the active device.

## measuring cross-talk or isolation

A coaxial antenna change-over relay or switch should

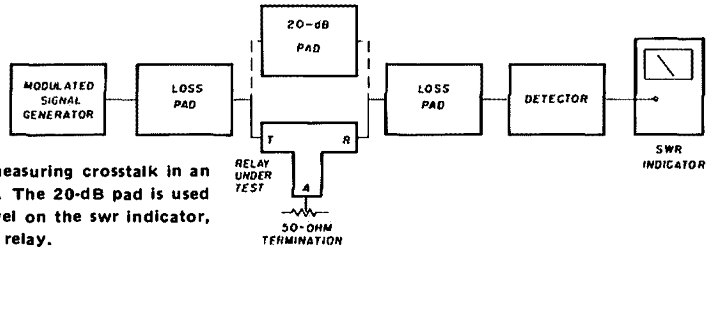



fig. 2. Arrangement for measuring crosstalk in an antenna change-over relay. The 20-dB pad is used to establish a reference level on the swr indicator, and is then replaced by the relay.

table 1. Reflection coefficient and VSWR vs return loss

| return loss (dB) | reflection coefficient | vswr | return loss (dB) | reflection coefficient | vswr |
|---|---|---|---|---|---|
| 1 | .8913 | 17.3910 | 31 | .0282 | 1.0580 |
| 2 | .7943 | 8.7242 | 32 | .0251 | 1.0515 |
| 3 | .7079 | 5.8480 | 33 | .0224 | 1.0458 |
| 4 | .6310 | 4.4194 | 34 | .0200 | 1.0407 |
| 5 | .5623 | 3.5698 | 35 | .0178 | 1.0362 |
| 6 | .5012 | 3.0095 | 36 | .0158 | 1.0322 |
| 7 | .4467 | 2.6146 | 37 | .0141 | 1.0287 |
| 8 | .3981 | 2.3229 | 38 | .0126 | 1.0255 |
| 9 | .3548 | 2.0999 | 39 | .0112 | 1.0227 |
| 10 | .3162 | 1.9250 | 40 | .0100 | 1.0202 |
| 11 | .2818 | 1.7849 | 41 | .0089 | 1.0180 |
| 12 | .2512 | 1.6709 | 42 | .0079 | 1.0160 |
| 13 | .2239 | 1.5769 | 43 | .0071 | 1.0143 |
| 14 | .1995 | 1.4985 | 44 | .0063 | 1.0127 |
| 15 | .1778 | 1.4326 | 45 | .0056 | 1.0113 |
| 16 | .1585 | 1.3767 | 46 | .0050 | 1.0101 |
| 17 | .1413 | 1.3290 | 47 | .0045 | 1.0090 |
| 18 | .1259 | 1.2880 | 48 | .0040 | 1.0080 |
| 19 | .1122 | 1.2528 | 49 | .0035 | 1.0071 |
| 20 | .1000 | 1.2222 | 50 | .0032 | 1.0063 |
| 21 | .0891 | 1.1957 | 51 | .0028 | 1.0057 |
| 22 | .0794 | 1.1726 | 52 | .0025 | 1.0050 |
| 23 | .0708 | 1.1524 | 53 | .0022 | 1.0045 |
| 24 | .0631 | 1.1347 | 54 | .0020 | 1.0040 |
| 25 | .0562 | 1.1192 | 55 | .0018 | 1.0036 |
| 26 | .0501 | 1.1055 | 56 | .0016 | 1.0032 |
| 27 | .0447 | 1.0935 | 57 | .0014 | 1.0028 |
| 28 | .0398 | 1.0829 | 58 | .0013 | 1.0025 |
| 29 | .0355 | 1.0736 | 59 | .0011 | 1.0022 |
| 30 | .0316 | 1.0653 | 60 | .0010 | 1.0020 |

provide at least 50 dB isolation between the transmit and receive connectors, but many do not, especially at uhf. Inadequate isolation can easily result in the destruction of transistors in the receiver when used in proximity to a high-power transmitter.

You can play it safe and measure the cross-talk, or isolation, of your change-over device by means of (you guessed it) the swr indicator. Since you hope to find the cross-talk down by 50-dB, which is well beyond the square-law region of the detector, a substitution method of measurement is employed.

Fig. 2 shows the test set-up for such measurements. Initially, a 20-dB pad is inserted as shown, and a 0-dB reference level established on the swr indicator. Since we hope to find the cross-talk between the transmit and receive ports down by well over 20 dB, the reference level should be set on the 40-dB range. Then remove the 20-dB pad and connect the relay as shown. The isolation will be equal to the change indicated by the swr indicator plus 20 dB.

## measuring vswr with a directional coupler

This one is for the vhf and uhf enthusiast, since directional couplers are usually used at frequencies of 200 MHz and higher. And although amateurs who are fortunate enough to have a directional coupler normally use it only for power measurements, it can be utilized to measure vswr by return loss.

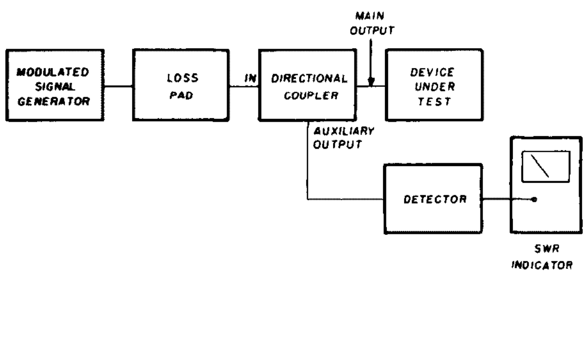
The equipment set-up for this measurement is shown in fig. 3, and the procedure is as follows:

1. Establish a reference level of 0 dB on the least sensitive range (but not less than 30 dB) commensurate with the available signal-source power.

2. Reverse the input and output connections to the directional coupler, keeping the detector on the auxiliary-output port.

3. Determine the return loss ($\alpha$) in dB from the reference level, as measured by the swr indicator.

fig. 3. Equipment configuration for measuring reflection coefficient and vswr by the return-loss method. Measurements are essentially independent of the coupling coefficient, but accuracy is severely limited by the directivity of the coupler, as explained in the text.



4. Determine the vswr from **table 1**, which shows the tabulated results of the following equations, where $\rho$ is the reflection coefficient.

$$p = 10^{\frac{-\alpha}{20}}$$

and

$$VSWR = \frac{1+\rho}{1-\rho}$$

Theoretically, measuring vswr by the return-loss method is accurate and quite simple to perform. In practice, however, the directional coupler can introduce large errors if its directivity is not substantially greater than the measured return loss. A good rule of thumb is to use a coupler whose directivity is at least 15 dB greater than the return loss. Since commonly available directional couplers generally have a specified directivity of only 20 or 30 dB, this would appear to limit accurate measurements to return losses of 5 to 15 dB, with the lower of these two values corresponding to a vswr of 3.57 (see table 1). However, even though the difference between the measured return loss and the coupler directivity is less than 15 dB, the technique is still usable for adjusting the equipment under test for minimum vswr.

In addition to coupler directivity, there are other factors which affect measurement accuracy, but these are of lesser importance. A complete discussion of measurement errors using directional couplers is beyond the scope of this article, but appears in reference 4.

I hope that this article has demonstrated the versatility of the swr indicator in the ham shack. Although the various commercial models mentioned earlier are generally available in limited quantities, there is no sure way of knowing that the supply will continue to exist. Even if you cannot find an instrument on the surplus market, it is a relatively easy unit to build, and will be just as accurate as the commercial versions. An equivalent, though slightly less complicated, instrument is completely described in reference 2.

Despite the length of this article, the primary *raison d'etre* of the swr indicator, that of use with the slotted line, has not been covered. This will be the subject of a future article.

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3. Joe Reisert, W1JAA, "VHF/UHF Antenna Matching Techniques," *ham radio*, July, 1976, page 50.
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5. *Operating and Service Manual, 415B Standing Wave Indicator*, Hewlett-Packard Company, Palo Alto, California 94304.
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**ham radio**

MICROPROCESSORS

# microcomputer interfacing: the vectored interrupt

This month we continue our discussion of computer interrupts, with emphasis upon the vector-interrupt hardware and software associated with the 8080A microprocessor chip. The three signals that are used in vector-interrupt circuits include INT (input pin 14 on the 8080A chip), INTE (output pin 16), and $\overline{\text{INTA}}$ which is not available on the 8080A chip but derived with external logic.

A positive clock pulse, from an interrupting device, supplies a logic 1 to the INT (*interrupt request*) input. This pulse generates an interrupt request which the CPU recognizes either at the end of the current instruction being executed or while the CPU is in the halt state. The

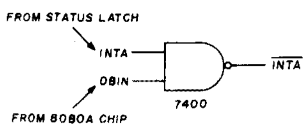


**fig. 1. The $\overline{\text{INTA}}$ control signal is generated from the DBIN and INTA control signals, only one of which is available on the 8080A chip.**

INTE, *interrupt enable*, output pin indicates the logic state of the interrupt-enable flip-flop within the 8080A chip. This internal flip-flop can be set (enabled) or cleared (disabled) with the aid of the 8080A microcomputer instructions:

| | | |
|---|---|---|
| **363** | DI | Disable interrupt flip-flop |
| **373** | EI | Enable interrupt flip-flop |

When cleared, the interrupt-enable flip-flop inhibits interrupts from being accepted by the CPU. The flip-flop is automatically cleared when an interrupt is accepted, and also by the RESET input signal applied at pin 12 of the 8080A chip.

The $\overline{\text{INTA}}$ or *interrupt acknowledge* control signal is


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generated by applying the INTA and DBIN control signals to a two-input NAND gate (fig. 1). A logic 1 at DBIN, *data bus in*, (pin 17 on the 8080A chip) indicates to external devices that the data bus is in the input mode. INTA is a positive clock pulse that is generated as a status output with the aid of a latch connected to the 8080A microprocessor chip.[1,2] We shall talk about the

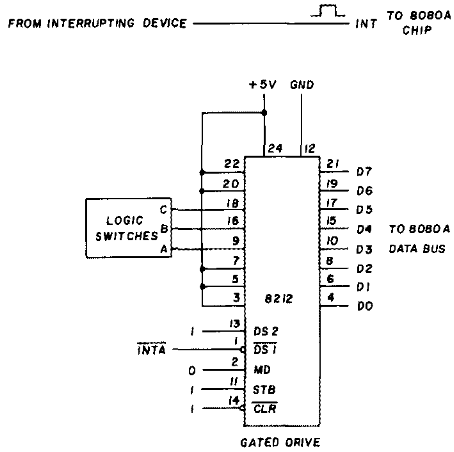


**fig. 2. Interface circuit for the jamming of a single-byte instruction into the instruction register of an 8080A microprocessor chip.**

status latch in a subsequent column. The interesting aspect of the $\overline{\text{INTA}}$ control signal is that it permits you to "jam" an interrupt-vector instruction byte directly into the instruction register in the 8080A chip. This can only be done during an interrupt, but nevertheless it is a unique and highly interesting operation that is possible with the 8080A microprocessor.

A simple circuit that demonstrates how a single-byte instruction can be jammed into the instruction register is shown in fig. 2. Assuming that the interrupt-enable flip-flop has been previously enabled by the instruction, 373,

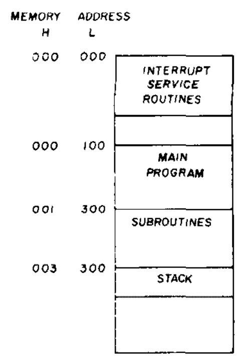


**fig. 3. Possible memory map for 8080A microcomputers.**

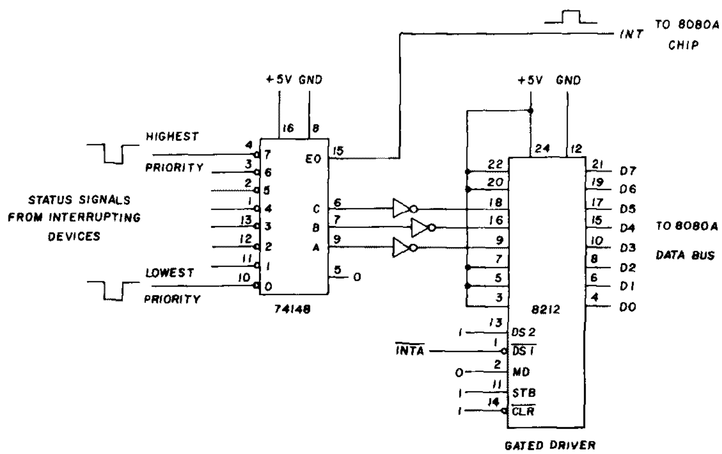


**fig. 4. Priority interrupt encoding scheme for an 8080A microcomputer.**

the interrupting device must supply a logic 1 input at INT in order to generate an interrupt request. The microcomputer finishes the current instruction and than generates the interrupt acknowledge signal, $\overline{\text{INTA}}$, which jams the desired vector instruction byte on the data bus and into the instruction register. Although any instruction byte can be jammed into the instruction register during an interrupt, usually the following eight instructions produce a useful result:

| | | |
|---|---|---|
| **307** | RST 0 | Call the subroutine that starts at HI = 000 and LO = 000 |
| **317** | RST 1 | Call the subroutine that starts at HI = 000 and LO = 010 |
| **327** | RST 2 | Call the subroutine that starts at HI = 000 and LO = 020 |
| **337** | RST 3 | Call the subroutine that starts at HI = 000 and LO = 030 |
| **347** | RST 4 | Call the subroutine that starts at HI = 000 and LO = 040 |
| **357** | RST 5 | Call the subroutine that starts at HI = 000 and LO = 050 |
| **367** | RST 6 | Call the subroutine that starts at HI = 000 and LO = 060 |
| **377** | RST 7 | Call the subroutine that starts at HI = 000 and LO = 070 |

Thus, the first sixty-four memory locations are reserved for *interrupt service routines* or *pointers*, extremely short programs, often consisting of only a single jump instruction. They tell the microcomputer what to do or where to go for a specified interrupt condition. Such routines precede the main program and associated subroutines in memory (**fig. 3**). If interrupts or restart instructions are not used, this portion of memory does not have any special significance.

**Fig. 4** is probably the simplest priority-encoder interrupt circuit that can be used with an 8080A microcomputer. The Intel 8212 chip is used as an 8-bit three-state buffer that inputs the instruction byte into the instruction register. The 74148 8-line to 3-line priority-encoder chip has the following truth table:

| inputs | | | | | | | | outputs | | | |
|---|---|---|---|---|---|---|---|---|---|---|---|
| **0** | **1** | **2** | **3** | **4** | **5** | **6** | **7** | **C** | **B** | **A** | **EO** |
| X | X | X | X | X | X | X | 0 | 0 | 0 | 0 | 1 |
| X | X | X | X | X | X | 0 | 1 | 0 | 0 | 1 | 1 |
| X | X | X | X | X | 0 | 1 | 1 | 0 | 1 | 0 | 1 |
| X | X | X | X | 0 | 1 | 1 | 1 | 0 | 1 | 1 | 1 |
| X | X | X | 0 | 1 | 1 | 1 | 1 | 1 | 0 | 0 | 1 |
| X | X | 0 | 1 | 1 | 1 | 1 | 1 | 1 | 0 | 1 | 1 |
| X | 0 | 1 | 1 | 1 | 1 | 1 | 1 | 1 | 1 | 0 | 1 |
| 0 | 1 | 1 | 1 | 1 | 1 | 1 | 1 | 1 | 1 | 1 | 1 |
| 1 | 1 | 1 | 1 | 1 | 1 | 1 | 1 | 1 | 1 | 1 | 0 |

The letter X means that the logic state is irrelevant.

The purpose of the circuit of **fig. 4** is to input the restart instruction, 3Y7, into the microcomputer. Five of the eight inputs to the 8212 chip are tied to a logic 1 state. The remaining three bits supply the encoded-vector address of the restart subroutine. By virtue of its truth table, the 74148 priority encoder chip provides eight priority levels. The inputs to this chip should be latched. The chip provides the three-bit binary output that corresponds to the highest valued priority input which is at logic 0 state. The inverters supply the three-bit Y component of the restart instruction. If there is a logic 0 at any of the inputs to the 74148 chip, a logic 1 output will be generated at the EO output (pin 15). This output serves as the input to the interrupt request pin, INT, on the 8080A chip. Upon receiving an interrupt request, the microcomputer responds with an interrupt acknowledge output, $\overline{\text{INTA}}$, that strobes the selected highest priority restart instruction into the instruction register.



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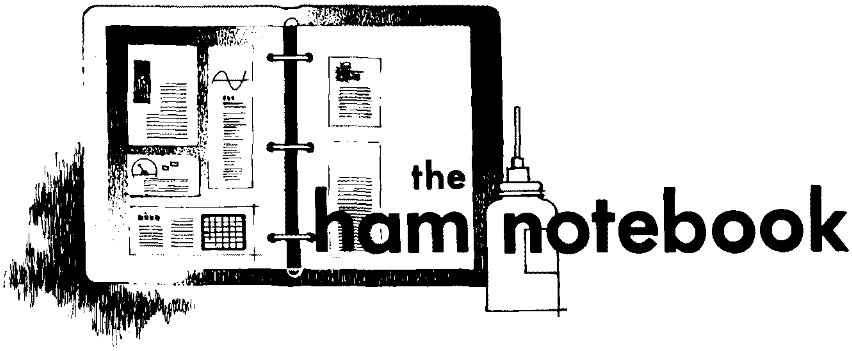



## WWV on the Heath SB-102

The SB-102 (and others of the Heath series) may be used to copy WWV on 5, 10, or 15 MHz by using the plug-in heterodyne crystal modification.[1] The bandswitch position and frequency of the heterodyne crystal depend upon the WWV frequency to be monitored. For instance, I use a junk-box crystal with a marked frequency of 8.001 MHz in conjunction with a homebrewed remote vfo operating above 5.6 MHz. This permits me to copy 15 MHz WWV while on the 14 MHz bandswitch position with the crystal tripling. Some crystals I tried would not oscillate properly on their third or fourth overtone in the SB-102 circuit (most did) and if a specific frequency is desired, a proper crystal should be purchased. However, the crystal available in your junk box should at least be given a trial.

Some combinations of crystal/LMO/-WWV frequencies in the 3.5 MHz bandswitch position are shown in table 1.

**table 1. crystal frequency chart for copying WWV on the Heath SB-102.**

| band | WWV frequency | injection frequency (upper/lower limits) (MHz) |
|---|---|---|
| 3.5 | 5.0 | 13.3936 to 13.8964 |
| 7.0 | 10.0 | 18.3936 to 18.8964 |
| 14.0 | 15.0 | 23.3936 to 23.8964 |
| 21.0 | 20.0 | 28.3964 to 28.8964 |

1. Injection frequency = LMO + WWV + 3.3964 (for compatible usb/CW)
2. Injection frequency = LMO + WWV + 3.3936 (for lsb)

This, in conjunction with the formulae, should be sufficient to calculate the needed/desired heterodyne crystal frequency. Remember, the LMO is at its lowest frequency (5.0 MHz) at the upper band edge and vice-versa; it always tunes backwards. As shown in table 1, for WWV at 5 MHz any crystal between 13.3936 and 13.8964 MHz will permit copy at some point between 3.5 and 4.0 MHz. Since the LMO actually tunes a little above and below the 5.0-5.5 MHz range, some margin is provided in case of error or crystal tolerance.

HC-6/U or HC-17/U crystals should be used. FT-243 types won't work. Also, the frequency of the heterodyne crystal should be reasonably close to that of the heterodyne crystal normally in use on the band in question. Check the SB-102 manual for other crystal information.

**Paul K. Pagel, K1KXA**

## TVI cure for the Kenwood TS-520

A couple of local amateurs were having TVI problems on channel 5 (New York) on two separate cable television systems. On other channels, with an outside antenna, there was also some TVI, but not as bad. I was asked to take a look at the problem and to recommend a solution. From the first, it looked like harmonic interference and not fundamental overload. Initially, I thought that rf from the Kenwood was getting into the ac line via the chassis and ac line cord, but after some filtering, additional bypassing and shielding on the chassis, I concluded that the problem was not there, since these measures did not help appreciably.

Next, I checked neutralization of the 2001A final amplifier tubes, but found them rock-solid from 80 through 10 meters.

Finally, I looked inside the final rf amplifier compartment and discovered the problem. I discovered that the final plate tuning capacitor had only one supporting plate and that the stator and rotor plates are unsupported at the front end. The rotor is grounded to the chassis by a three-legged spring clip that helps to hold the rotor in line. The rotor shaft is attached to a fiber shaft by a flexible coupling, and the fiber shaft exits the final amplifier compartment through a metal sleeve or collar.

Harmonic rf energy is radiated through the sleeve, flows over the outside of the compartment, and thence to the ac line. I also noticed that the fiber shaft is attached to a metal shaft by a small pin, and that the metal shaft makes a dandy little antenna to radiate the unwanted harmonic.

To cure the problem in your TS-520, proceed in the following manner:

1. Remove the three screws that hold the plate tuning capacitor to the wall of the final amplifier compartment.

2. Make a cover of thin sheet aluminum large enough to cover the three screw holes so it can be fastened to the compartment wall.

3. Clean the area of contact so that it makes a good rf-tight joint and drill a hole in the aluminum plate just large enough to pass the shaft of the variable capacitor. File a small notch in the aluminum at the edge of the drilled hole to pass a grounding wire (this wire will be soldered to the above-mentioned sleeve).

4. Scrape the sleeve where the wire will be soldered to it, for a solid joint, and solder one end of a short length of wire to it.

5. Disconnect the flexible coupling from the shaft of the variable capacitor.

6. Immediately behind the front panel of the TS-520 where the variable tuning capacitor shaft passes through, locate a locking spring clip; loosen the bolt and pull the shaft out an inch or so, so that you can slip the aluminum plate over the shaft and the wire soldered to the sleeve.

7. Replace the screws in the capacitor mounting plate and place a solder lug at the upper right-hand screw (looking from the front of the rig back at the plate). Tighten the screws and solder the free end of the wire to the lug.

You will also notice a white wire passing through a rather large hole in one wall of the compartment. Try to fashion the aluminum cover plate in such a way as to cover this hole, leaving room only for the wire.

The "fix" described here will remove interference from the *picture* carrier, but will not solve the sound carrier interference. To get rid of the latter, proceed as follows:

1. Buy, beg or steal a sheet of perforated aluminum of the kind that has small holes and is available at most hardware stores. Do not get the slotted variety!

2. Cut a piece of this screening material large enough to cover the slots in the rf final amplifier compartment. A piece about 4½x5 inches (11.4x12.7cm) should do the trick. The slots are to permit the blower to pull cooling air through the compartment, and the screen cover you install will not appreciably alter this flow of cooling air. At the screw holes, brush the compartment, walls and the aluminum screening to make a good rf-tight contact.

3. Mount the "shield" made in this manner to the side of the rf amplifier compartment with sheet metal screws or other hardware.

4. On top of the rf amplifier compartment, fit a small piece of screening material just large enough to cover the two holes, and attach it by means of the screw that is close by.

5. The bottom of the TS-520 cabinet has a series of various-sized holes used to pass alignment tools for tuning the different circuits. These also pass rf and should be covered for a complete shielding job.

6. Remove the four feet and the center screw from the cabinet and cut a piece of aluminum screen large enough to cover all of the "tuning" holes and *fasten it in place on the* inside cabinet floor with the center screw and the four feet.

After completing these operations, the only way rf can escape the transmitter is through the center conductor of the coaxial cable — just where you want it!

I verified the TVI cure by running the transmitter and an SB-220 at full rated output next to an old TV set that had a barely discernible picture and poor sound. Oh yes, by the way, I used an open-wire feed line to my antenna. No interference on any channel! Needless to say, my amateur buddies, the *cable company and the subscribers are* very happy; because when you get into the cable you get into the city!

Stacky Stackhouse, W3FUN

## simple tune-up for Drake gear

I have found a very simple method of tuning my Drake T4XB, R4B transmitter-receiver combination that certainly will work with other models of the T4 and R4, and may even work with other equipment. The advantage of my tuning method is that it takes less time, it is easier on the final amplifier tubes and little or no carrier appears on the operating frequency during 95% of the procedure. Here's how it is done:

1. Set the receiver to the desired operating frequency.

2. Set the transmitter for CW mode operation, place the transceive switch in the "spot" position and tune the transmitter vfo until the spot signal is heard in the receiver. The transmitter "gain" control may have to be advanced slightly, perhaps even one-quarter turn, to obtain enough signal to be useful.

3. Now, using the receiver's S meter, tune both the transmitter and receiver preselectors for maximum S meter reading.

4. While watching the S meter, vary both the transmitter "plate" and "load" tuning controls for maximum signal, and repeat this step until no further signal increase is possible.

5. Turn the transceive switch to "separate" (with either receiver or transmitter controlling the frequency, as desired) and finish the tune-up according to the instruction book.

J. L. Kofron, W7DIM

## etch tank

If you etch printed circuit boards regularly, the professionally styled etch tank shown in fig. 1 works very well. It features an aerator that allows you to bubble air from a fish tank pump through the holes in aerator. This speeds up the etching process.

The basic material for the tank is 1/8 inch (13mm) plexiglass. Needless to say, the joints must be water tight.

Gary L. Tater, W3HUC

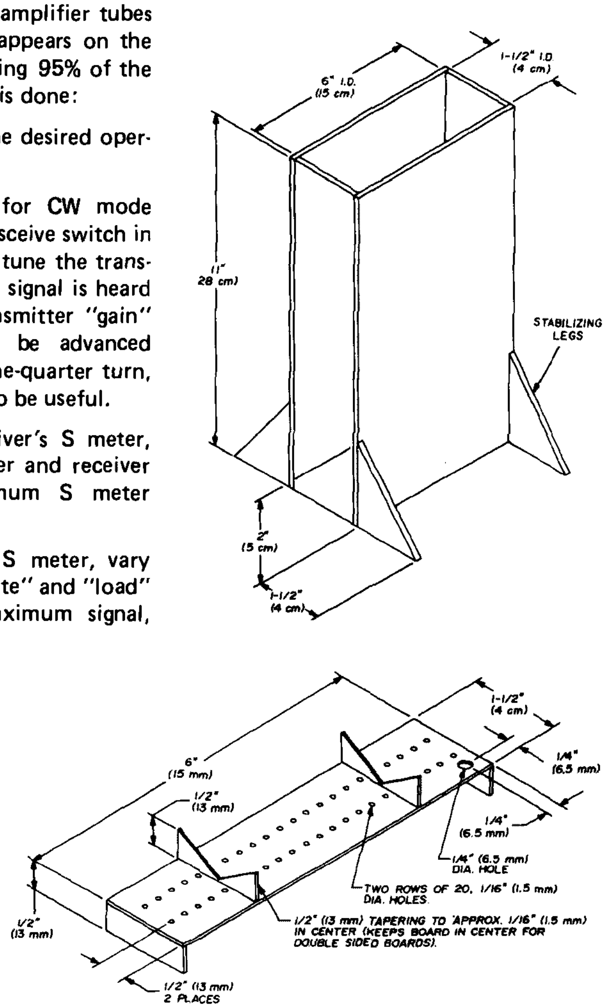
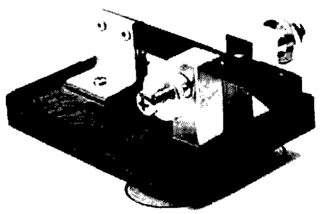



**fig. 1. Professionally styled etch tank for printed-circuit boards uses aeration to speed up the etching process. Aerator sits on bottom of the tank; air is furnished by a fish tank type air pump. All material is 1/8" (13mm) plexiglass or lucite. All joints must be water tight.**

# NEW products

## side swiper morse keyer



The Swedish firm, Kungsimport, has introduced its new Side Swiper, a keying lever designed by SM6CKU, for use by amateurs or commercial operators throughout the world. The Side Swiper may be used by itself to produce dots and dashes in an easy and natural manner or it may be used as a keying lever for an electronic keyer.

Easy and fun to handle, the Side Swiper requires only a few hours familiarization and provides a distinctive personal fist or the characteristic swing of an old-time operator. CW produced by the Side Swiper is easily recognizable and lacks the monotonous, mechanical-sounding CW of the electronic keyer or bug.

The Side Swiper is personally and individually hand crafted with a wood base and metal armature and contacts, and is priced at $13.00, airmail shipment included. Your name or callsign may be engraved on the wooden base for an extra dollar. For additional information write Kungsimport, Post Office Box 257, S43401 Kungsbacka, Sweden, or use *check-off* on page 126.

## Amphenol dummy load coaxial connector

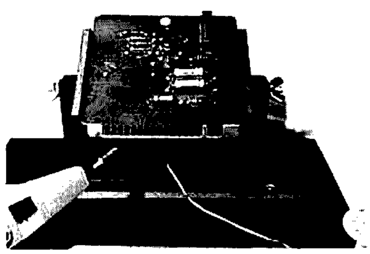
A new combined illuminated power output indicator and dummy load that provides a visual indication of low-power transmitter performance was recently introduced by Bunker Ramo RF Division.

The new device, designated Amphenol® Model 83-888, consists of a light-terminated coaxial load mounted in a modified uhf connector (PL 259-type) body. It can be used to tune your low-power transmitter for maximum output — the brighter the light, the greater the output.

Amplitude modulation can also be checked quickly, because the brilliance of the light increases and varies when someone talks into the microphone when the transmitter is functioning properly. The 83-888 can also be used to check ssb carrier balance: there will be no glow if the carrier is correctly nulled, but the light will glow with modulation.

The new power output indicator mates with standard uhf receptacles (5/8 - 24 thread). Its nominal impedance is 52 ohms, the frequency range is 0-30 MHz, power rating is 4 watts, maximum, and VSWR is 1.1:1 at 27 MHz. For additional information, contact Bunker Ramo RF Division, 33 East Franklin Street, Danbury, Connecticut 06810 or use *check-off* on page 126.

## printed-circuit board holding fixture



The W.N. Wellman Company, has introduced a versatile holding fixture to facilitate PC board parts-mounting and soldering. The fixture is adjustable to accomodate most PC board sizes and may be positioned to provide convenient and easy assess to either surface of the board. The base of the holding fixture is provided with resilient feet to hold it firmly on a flat work table or bench.

The price is $7.95 postpaid within the continental U.S. Missouri residents please add $.25 tax. For additional information, write the W.N. Wellman Company, 451 Saline Road, Post Office Box 722, Fenton, Missouri 63026 or use *check-off* on page 126.

## Larsen adds TLM trunk lid mount mobile antennas

An all-new mobile antenna mounting option has been added to the wide variety of mobile mounts available from Larsen Electronics. The TLM stands for Trunk Lid Mount, and it is available in seven different variations so that every antenna and mounting condition can be met.

This new Larsen trunk lid antenna mount follows the configuration of the popular Larsen magnetic mount, and results in a low-silhouette installation. It is finished in highly-polished chrome and is provided with a special molded gasket to protect the car finish. The mount clamps to the vehicle with stainless-steel, hollow-head screws that provide a positive ground and will not rust or "freeze" in place. The mount comes factory- assembled with 17 feet of RG58A/U coaxial cable and plug. The TLM is available with hardware to accomodate all standard antenna mounts including Motorola, ASP, GE, RCA and — of course — Larsen. It may

be purchased complete with antenna, or without either antenna or mounting hardware.

For additional information, write Jim Larsen, Larsen Electronics, Box 1686, Vancouver, Washington 98663; telephone (206) 573-2722 or use *check-off* on page 126.

## dual dc power supplies

Mid-Continent Communications Company recently introduced a new power supply package featuring two independent power supplies, having isolated outputs, contained within a single enclosure. The Mcc/103 power supply helps eliminate bench clutter when designing with ± 15 Vdc and is conveniently dimensioned to facilitate stacking with identical units or with other items of equipment. The wood-grained aluminum enclosure measures 2.25 inches high, 10 inches wide and 5.7 inches deep (5.7x25.4x14.5cm). A backlighted and flush-mounted panel meter permits accurate monitoring of current and voltage for either supply with the desired functions selected by a meter mode switch.

Variable voltage and current limit controls allow easy setting of the voltage to typical values between zero and 25 volts, and the current to typical values between 10 and 250 milliamps. Since the two supplies are completely independent, it is possible to connect them in series or parallel to provide, typically, 50 volts at 250 milliamps or 25 volts at 500 milliamps. A unique feature of the Mcc103 is the provision of two LEDs (light-emitting diodes), one for each section, to indicate when the output current exceeds a preset limit point. This feature can help prevent confusion and false indications during circuit design and testing.

The Mcc103 employs two voltage regulator boards designed for excellent regulation and temperature stability. All components, including the rectifier and filter, are mounted on the printed-circuit board and the boards themselves can be lifted from the enclosure without unsoldering any wiring, permitting easy removal for maintenance and repair. Internally mounted *Thermatab* transistors are used, eliminating shock or shorting hazards. The current-limit system and the internally mounted ac fuse provide total short circuit protection for both the power supply and the load.

Power requirements are 105 to 125 Vac at 25 watts, and the line regulation is ± 0.2 percent maximum. Load regulation is ± 0.05 volt, maximum, while ripple and noise are 1 millivolt rms, maximum. The Mcc103 is priced at $99.95 plus $2.50 postage and handling. Missouri residents please add 3 percent sales tax. For additional information, write to Mid-Continent Communications Company, Box 4407, Kansas City, Missouri 64127, or use *check-off* on page 126.

## H-P offers digital multimeter with touch-hold reading probe

The Model 3435A, 3-1/2 digit multimeter from Hewlett-Packard has a unique "touch-hold" probe available as an accessory. It lets the user "freeze" the reading on the display — a convenience when probing closely-packed circuit boards. Accurate enough for both bench and field use, the new digital multimeter is autoranging on ac and dc volts and resistance. Ac and dc current ranges are selected manually. Lighted front panel annunciators display the function and its units.

The digital multimeter covers a dc measurement range from 200 mV full scale to 1200 V full scale with a mid-range accuracy of ±(0.1% of reading + 1 digit). Ac measurement range is 200 mV full scale to 1200 volts rms full scale with a mid-range accuracy of ±(0.3% of reading + 3 digits) over a 30 Hz to 100 kHz bandwidth. Ac and dc current measurement range is from 200 microamps to two amps. Dc current accuracy for the 200 $\mu$A to 20 mA range is ±(0.3% of reading + 2 digits). Ac current measurements are made over a frequency band of 30 Hz to 10 kHz with a mid-band accuracy of ±(1.7% of reading + 4 digits). Resistance range is 10 milliohms to 20 megohms with a mid-range accuracy of ±(0.2% of reading + 2 digits). Open circuit voltage on the ohms terminal when set to its lowest range does not exceed 5 volts, preventing damage to most solid-state devices.

Input protection is provided to 1.2 kV on any dc range, 1700 V (dc + peak ac) on any ac range, and 250 V rms on any resistance range. A front panel fuse protects the instrument from overload when measuring current.

A choice of two power supplies is offered: internal ac power supply and rechargeable lead acid batteries. The standard 3435A Digital Multimeter comes with an internal ac power supply and rechargeable lead acid batteries. Option 001 is a lightweight portable case, ac line power only. Option 002 is the ac line power only with a rack and stack case.

U.S. price of the standard Hewlett-Packard Model 3435A digital multimeter with rechargeable lead acid batteries is $400. Option 001, ac only in custom plastic case is $335; 002, ac



More Details? CHECK—OFF Page 126

only, in a rack and stack case is $365. Delivery is from stock. "Touch-Hold" reading probe U.S. price is $40, RF Probe is $87. High Voltage Probe is $75.

For additional information, write Inquiries Manager, Hewlett-Packard Company, 1501 Page Mill Road, Palo Alto, California 94304; telephone (415) 493-1501, or use *check-off* on page 126.

## six new calculators, video game and digital watches from National Semiconductor

National Semiconductor, Consumer Products Division, recently announced 6 new calculator products, including 3 advanced slide rules bearing the National Semiconductor brand name, an ideal family-type model, a teaching version of the QuizKid, a complete line of men's and ladies' LED digital watches and a video game series bearing the name Adversary.

The model 852 calculator offers scientific notation or floating decimal point system with reformating capability from one system to the other. The 852 features algebraic logic, two level parentheses, trig and log functions, degree/radian conversion. It offers an 8-digit Mantissa in floating point system and 5-digit Mantissa plus 2-digit exponent in scientific notation, plus many other features. Suggested retail, $29.95.

The model 4650 calculator offers algebraic logic, two level parentheses, full accumulating memory addressable in all 4 arithmetic functions, log and trig functions, degree/radian conversion, rectangular/polar coordinates, 8 Mantissa digits, 2-digit exponent display, plus many other features. Suggested retail, $59.95.

The model 4660 calculator that displays 10-digit Mantissa in floating point system and 10-digit Mantissa plus 2-digit exponent in scientific notation, with algebraic logic, two level parentheses, three separate, addressable, accumulating memories, trig and log functions, decimal degrees and degrees, minutes and seconds conversions, polar and rectangular coordinate conversion and many other features. Suggested retail, $79.95.

The QuizKid II allows the child to see a timed series of 10 problems which appear automatically in the display. The

child is required to key in the answer to the problem shown in the display. Over 1200 problems are automatically generated by the calculator, permits selection of the type of arithmetic function (i.e., addition, subtraction, etc.) which is displayed automatically in random sequence, and offers a slow/fast speed control key to adjust time allowed for child to enter answer. Suggested retail, $24.95. Also available is an optional game adapter which connects two Quiz-Kid II's for a contest.

The QuizKid III – retains all the abilities of QuizKid II plus games for over 6,500 additional problems. Contains amateur and pro keys for adjusting complexity of problems, and a complex key for problems to be automatically displayed with one of the factors missing but with the answer given. This model is being test marketed.

The video game – Adversary – features a choice of 3 games: tennis, played by two players on a green court; ice hockey, played by one or two players on blue ice; and handball, played by two players on a brown court. All games are in full color, have realistic sound effects when the ball or puck strikes a surface, and offers a choice of 3 individually selectable paddle sizes. Serves are controlled by players, not by random, and scoring is automatically displayed in large easy-to-read numbers after each point is scored. Individual controls enable players to sit in their favorite chairs to compete. The game offers 7 modes of operation: 3 modes of 2 players, 3 modes of a player against himself and one mode with a player against the machine. A special feature allows "time out" during play without changing the score. Suggested retail, $99.95.

The model 830 Datachecker is an ideal calculator for shopping, taxes, family finances, and features an 8-digit LED display and floating decimal, and is designed to operate like the shopper thinks: add and subtract items, subtotal items, efficiently check grocery bills, balance checkbook. (Battery not included). Suggested retail, $14.95.

For further information contact Georgene Berglund, Public Relations, National Semiconductor Corporation, 2900 Semiconductor Drive, Santa Clara, California 95051; telephone (408) 732-5000, or use *check-off* on page 126.



More Details? CHECK–OFF Page 126

## Cushcraft builds new research and production facility

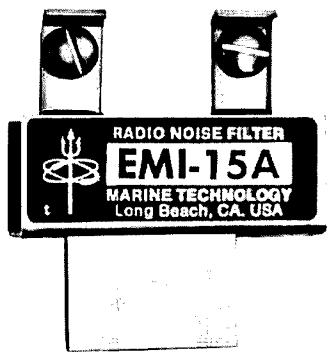
Construction is underway on the first phase of Cush Craft Corporation's new 50,000 square foot antenna research and production facility.

All manufacturing operations, executive offices and research were moved to the new facility at Grenier Industrial Park in Manchester, New Hampshire at the end of November 1976.

Bob Cushman, Cush Craft treasurer, reports this all new plant is designed especially for antenna manufacturing. It will allow increased production of current amateur, citizen's band and commercial antennas, plus the introduction of several new antenna types.

The new production lines and equipment have been planned for several years and, when fully operational, will be a model for the industry, allowing Cush Craft to maintain its traditionally high value standards.

## automotive ignition noise filter




Marine Technology recently announced its new EMI-15A filter designed to control ignition system generated noise in mobile two-way radios.

The introduction in 1975 of high-voltage, solid-state ignition systems in all American automobiles focused attention on the need for better suppression of ignition interference. Fully fifty per cent of engine noise problems in an automobile come from the ignition system.

The EMI-15A is a three-element low-pass LC filter designed to control igni-



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*tion system-generated noise.* It prevents coupling of ignition impulses into the vehicle primary wiring system and supplements suppressor-type spark plugs and wiring. Installation is made at the battery connection to the coil or electronic ignition system.

Suggested retail price is $6.95, and the units are available from stock. For additional information contact Morgan Cox, Marine Technology, 2780 Temple Avenue, Long Beach, California 90806; telephone (213) 427-6443, or use *check-off* on page 126.

## mobile radio case

If you worry about leaving your mobile radio in your car unattended, and if is is not convenient to carry your radio when you leave your car, Platt Luggage may have the solution to your problem. Platt has introduced a new line of rugged, lightweight, molded cases for mobile radio equipment and accessories. Molded from high-density polyethylene with a double wall construction, the cases incorporate shock-absorbent polyfoam interiors that can be custom-cut to fit nearly any size and shape of mobile equipment.

A Piatt case makes it convenient to carry your radio if you want to take it with you and it protects the radio if you want to leave it in the trunk of your car. The case is ruggedly handsome and is dent, shatter and scuff proof. Two sizes are available to fit nearly any requirement: CB-1 measures 12x8.5x4.25 inches (30.5x21.6x10.8cm) and CB-2 measures 13.5x10x4 inches (34.3x25.4x10.2cm). Finished in black, they make attractive companions for your equipment and are guaranteed for five years. The CB-1 sells for only $14.95 and the CB-2 is priced at

$19.95. For additional information, write to Platt Luggage, Incorporated, 2301 South Prairie Avenue, Chicago, Illinois 60616, or use *check-off* on page 126.

## oscilloscope and monitor adapter

I-f circuit waveform observations along with ssb and a-m transmitter signal monitoring has been made possible by Leader Instruments Corporation through the introduction of its new LBO-310 Ham Oscilloscope.

The new 3-inch (7.5cm) scope, which has a vertical sensitivity of 20 mV p-p division and a vertical bandwidth from dc through 4 MHz will also indicate tuned condition for RTTY operation as well as facilitate ssb signal observation through the use of an internal, two-tone test generator. The LBO-310, in combination with the LA-31 adapter will also provide continuous monitoring of rf output to 500 watts. Maximum input to the vertical amplifier is 600V, dc + ac peak-to-peak at a 1 megohm impedance. Transmitter monitoring is from 1.8 to 54 MHz at power levels from 2 to 500 watts with a deflection sensitivity of 1 watt per division into 50-ohm or 75-ohm impedances. The LBO-310 is priced at $269.95.

The LA-31 adapter which makes it possible to monitor the output waveform and power of both ssb and a-m transmissions from 5 to 500 watts over the frequency range from 1.8 to 54 MHz. It sells for $22.95.

For further details write Leader Instruments Corporation, 151 Dupont Street, Plainview, New York, 11803 or use *check-off* on page 126.



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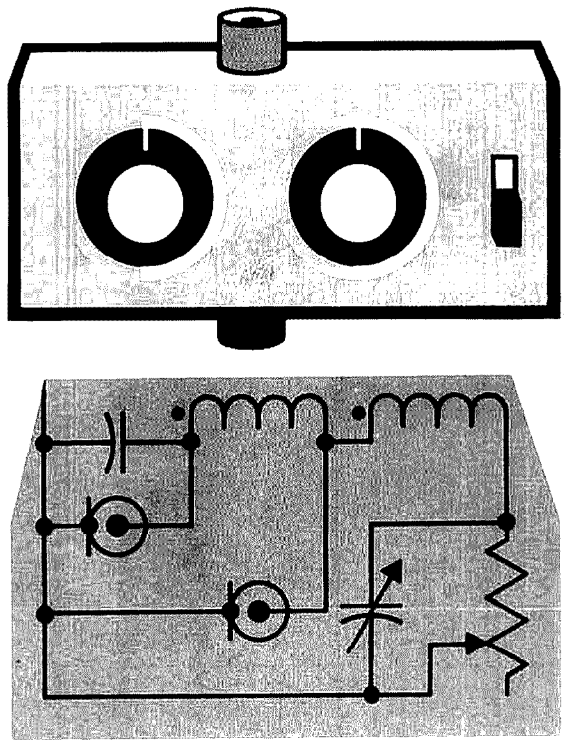
one dollar

# ham radio
magazine

## FEBRUARY 1977



# accurate rf noise bridge



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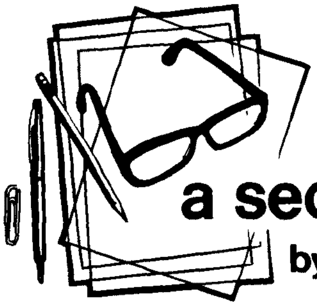


# a second look

**by Jim Fisk**

In its own way, Amateur Radio is starting to face one of the same problems that confronts everyone else in the world — pollution. We are squeezing more and more people into the same amount of frequency spectrum, and this is creating a congested mess for the Amateur operator. Mode and QSO battles rage every night with CW on top of RTTY, one net against another, and contester vs casual operator.

Some Amateurs have been looking for the 1979 World Administrative Radio Conference to alleviate congestion by expanding the Amateur bands, but as you can see in this month's presstop, it doesn't look promising. That means we've got to put modern technology to work to clean up our own act. We should be pushing for cleaner emissions from transmitters, and receivers that are able to cope with the overcrowding. However, the basic question arises, "How do you realistically compare various Amateur equipment?" Quite *frankly*, with things the way they *are* now, you *can't*.

Each manufacturer wants to sell his equipment with its own special features, and the published performance specifications naturally reflect *those* features. As a point of fact, the art of specification writing hasn't kept up with technology; gone are the days when receiver sensitivity and selectivity were the only things you looked for in a receiver — today's receivers must contend with an abundance of closely-spaced, strong signals, but receiver "specmanship" has changed little since the 1930s. The problem is compounded by the fact that manufacturers seldom use similar methods to test their products!

The problems of comparing transmitters is minimal because the published specifications are relatively straight forward: Essentially so many watts at some level of intermodulation distortion, during a two-zone test. This is a basic testing procedure that produces understandable, comparable results. However, there's still room for improvement. What is the impedance-matching range of the output tank circuit, for example, or the dynamic characteristics of the ALC circuit?

Receivers pose a much larger problem. The ads for Amateur receivers still reflect the age-old sensitivity/selectivity hangup — performance features that don't tell you how well the receiver is going to stand up under the high signal density of the real world. Some of today's more notable writers add further confusion by designing so-called "super" receivers and testing with different methods to different standards.

As it is, the problems cannot really be laid at the feet of any one group. Conversations with manufacturers indicate that there is a lack of authoritative information on the subject, so they are generally in a quandary when it comes to receiver testing. The commercial receiver manufacturers are faced with much the same problem, and many have turned to the standards established by the English Ministry of Posts and Telecommunications, or the German FTZ (their equivalent to the FCC). Recently some receiver standards have been set up for Marine equipment, but there's nothing available for the high-frequency bands.

In our studies of this problem, the cooperation of Amateur Radio manufacturers has been outstanding. One manufacturer even said that he would try to get any usable set of receiver test standards accepted by his company! Only one offered any resistance, and he refused to discuss the issue, apparently feeling that any disclosure of his testing methods would reveal proprietary information about his equipment.

What's needed by the Amateur Radio industry is a good set of methods and standards that will be followed by everyone. A recent article by W7ZOI (*QST*, July, 1975) lays down a good set of guidelines for checking the sensitivity, blocking, and IMD characteristics of a receiver, although the article has evidently been largely overlooked. With a little refinement to adequately cover all pieces of equipment, this would provide an excellent starting point. Obviously someone has to spearhead the effort to improve standards for Amateur equipment, and we at *ham radio magazine* are willing to serve as a clearinghouse for your ideas. If we can get our heads together, perhaps we can pound out a set of specifications that will make all of our lives easier when it comes time to buy new equipment. And in the long run, a realistic set of standards will inevitably result in better Amateur equipment for all of us.

Jim Fisk, W1DTY
editor-in-chief

# presstop

SUSPENSION OF FCC LICENSE FEES went into effect January 1st. The fee suspension applies to all FCC licenses — broadcast and commercial as well as Amateur and CB. It's the result of a Federal District Court decision stemming from a suit brought by cable TV and other interests who had charged that the FCC's fee structure was arbitrary and thus improper.

Duration Of The Suspension will probably depend on Congressional action to provide guidelines, and details are still being worked out. Suspension of license fees could favorably affect the growth rate of Amateur Radio if historic relationship between license fees and applications still hold. Amateurs who plan to apply for licenses or license renewals are requested not to send any fees with their applications.

AMATEUR FREQUENCY ALLOCATIONS PROPOSED in FCC's long-awaited Third Notice of Inquiry concerning the 1979 World Administrative Radio Conference don't provide us with the hoped for new HF bands but do expand some of the most important present bands. Perhaps the most significant change is the moving of several lower band edges down: two totally new allocations, one LF and the other in UHF, also offer intriguing possibilities.

A Band-By-Band Breakdown, shows the FCC plans to propose the following Amateur assignments:

| | |
|---|---|
| 1600 Meters: | 160-190 kHz (new, exclusive) |
| 160 Meters: | 1750-1800 kHz (new, shared), 1800-1900 kHz (exclusive), 1900-2000 kHz LOST |
| 80 Meters: | 3500-3900 kHz (exclusive), 3900-4000 kHz (shared) |
| 40 Meters: | 6950-7000 kHz (new, exclusive), 7000-7100 kHz (exclusive), 7100-7300 kHz (shared) |
| 20 Meters: | 13950-14000 kHz (new, exclusive), 14000-14350 kHz (exclusive), 14350-14400 kHz (new, exclusive) |
| 15 Meters: | 20700-21000 kHz (new, exclusive), 21000-21200 kHz (exclusive), 21200-21450 kHz LOST — but, encouraging informal discussions with Maritime people (who were to take the 21.200-21.400 MHz segment) suggest they may take 20.650-21.000 MHz instead, leaving 21.000-21.450 MHz exclusively Amateur, as it is now. |
| 10 Meters: | 28000-29700 kHz (exclusive, no change) |
| 6 Meters: | 50.0-54.0 MHz (exclusive, no change) |
| 2 Meters: | 144-148 MHz (exclusive, no change) |
| 1¼ Meters: | 220-225 MHz (shared, with radio-location and land mobile — read that Class E CB!) |
| 3/4 Meter: | 420-450 MHz (shared, with some change in priority) |
| 3/8 Meter: | 902-928 MHz (new, shared), 935-938 MHz (new, exclusive) |

Above 1 GHz, All The Bands we presently have are included in the proposal along with specific sub-bands for the Amateur Satellite Service. The Notice of Inquiry also proposed that the present microwave allocations, except for 48-50 GHz, be world wide to encourage international exchanges between Amateur researchers.

REQUESTS FOR MULTIPLE NOVICE exams must be accompanied by the name and necessary qualifications of the person (or persons) planning to administer the exams, names of the individuals to be examined and the date you expect to administer the exams. FCC reports they've been getting a number of exam requests such as "I'll need a dozen or so exams in the next six months" which they've had to reject.

PROPOSED FURTHER DE-REGULATION of the Amateur Radio Service has emerged from the FCC in the form of Docket 21033 which would revise Part 97 of the FCC Regulations to:

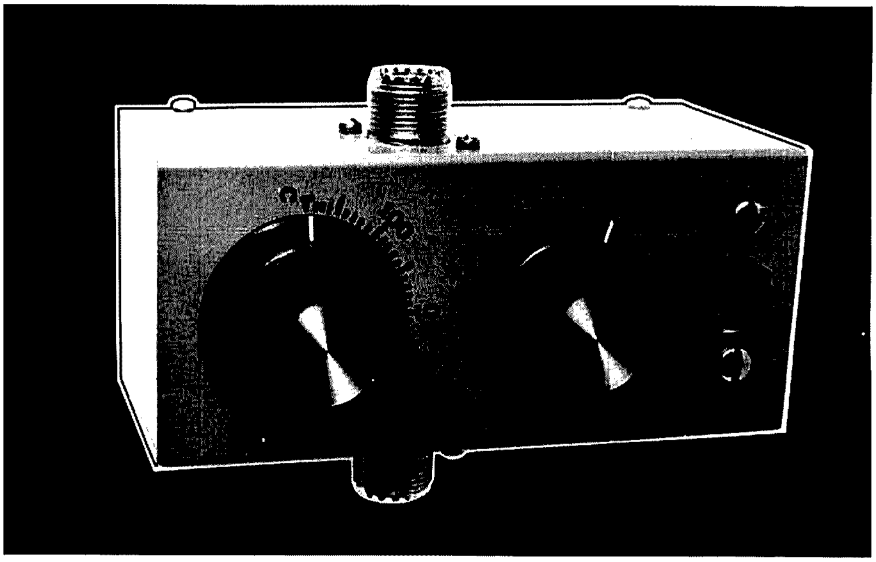
Permit repeater, auxiliary, and remote control operation of Amateur stations under Primary, Secondary, and Club station licenses, and to continue the issuance of separate licenses for such operations;

Delete the requirement that the transmissions of so-called "open" automatically-controlled repeater stations be recorded and the recordings be retained for a 30-day period, and making minor revisions to the logging requirements for remotely-controlled stations; and

Allow Amateur licensees greater flexibility in the choice of frequencies for repeater and auxiliary station use.

The Commission Also Invited comments on the adequacy or the like of current spectrum-management techniques in the Amateur Service. Comments are due by April 1, 1977 and reply comments are due by April 15, 1977.

AMSAT'S A-OD ORBIT is expected to have the following parameters if preliminary calculations are correct: Period — 102 minutes; Apogee — 915 km (567 miles); Perigee — 830 km (515 miles); and an angle of inclination to the equator of 99 degrees. The sun-synchronous satellite is expected to have a DX "range" of approximately 6440 km (4000 miles). LANDSAT C launch expected October, 1977.



# improvements to the RX noise bridge

## Circuit and construction improvements for this simple device improve accuracy and measurement range for impedance measurements from 3.5 to 30 MHz

How often have you wanted to know the exact impedance – both the resistive and reactive components – of your high-frequency antenna? Although the venerable swr meter recognizes the situation which makes the transmitter happy, it very often is misleading with regard to actual antenna resonance and related matching adjustments. Amateur literature has approached the subject of measuring complex impedance in the high-frequency range a number of times. In spite of all this, the use of these techniques by amateurs is still relatively uncommon. This is indeed unfortunate, since many antenna adjustments become systematic if resistance and reactance are accurately known.

The rf noise bridge is one of several devices that can be used to make this type of measurement. YA1GJM's excellent article originally got us interested in this device.[1] Subsequently, we have pursued modifications to the instrument which both improve its accuracy and extend its range. Ultimately, an rf noise bridge unit resulted which has the following significant characteristics:

1. Measurement accuracy of 3 ohms rms* for both the resistive and reactive components of representative rf loads.

A parts kit for the noise bridge is being made available in conjunction with this article. For information and prices write to G. R. Whitehouse & Co., 10 Newburg Drive, Amherst, New Hampshire 03031.

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2. Capability, from 3.5 to 30 MHz, to measure complex impedances equivalent to any point within a 5:1 swr circle on a Smith chart with a $Z_o = 50$ ohms.

3. A calibration concept which does not require laboratory standards and is valid over the entire frequency range up to 30 MHz.

4. Construction cost under $20.00.

With reasonable care, anyone can achieve these same results. All you have to do is follow the instructions on construction and calibration which are included in this article. In addition to these tips we will cover a wide range of other topics relative to the noise bridge. The first two sections describe the basic principles of the noise bridge and some history regarding its development. The next three sections describe our principal contributions — extending the range and improving the accuracy of the instrument. Details on how to build, checkout, calibrate, and use the improved instrument are concentrated in the last three sections.

## noise bridge description

An rf noise bridge may be used to measure complex impedances — normally a very difficult thing to do without laboratory instruments. It works as shown in the block diagram of **fig. 1**. An unknown impedance to be measured is connected to the input terminal of the bridge; a receiver is connected to the output terminal where it is used as a sensitive, frequency-selective detector. Contained within the bridge is a wideband noise source which provides a signal over the frequency range of interest. Reference devices with known rf impedances are used to balance the bridge section and to measure the unknown impedance.

A schematic diagram of the bridge portion is shown in **fig. 2**. It works in the following manner: Wideband noise is injected into two legs of the bridge in equal quantities via a core transformer, T1. With the unknown impedance connected and the detector (your receiver) set to the desired frequency, $R_p$ and $C_p$ are adjusted for the deepest obtainable null. There is some interaction between the two adjustments, depending upon frequency, so that a readjustment between the two controls may be required to obtain the deepest null. When the null condition is achieved, the value of the unknown impedance is equal to the parallel combination of $R_p$ and $C_p$. To permit measurement of both positive and negative values of parallel capacitance, the zero value of the $C_p$ dial is set with the variable capacitor, $C_p$, approximately half-meshed. To balance the bridge with this offset, a fixed capacitor, $C_f$, is placed across the unknown terminal. This forces the bridge to balance with a purely resistive load when the half-meshed $C_p$ is equal to $C_f$. However, by labeling the $C_p$ dial zero at this point, the correct capacitance is registered by the instrument.

Several advantages of the noise bridge concept now become apparent:

1. The frequency at which the measurement is being made is determined by the detector (your communications receiver), and this should be very accurate.

2. Measurement of inductive reactance does not require an accurate variable inductance which, in practice, is difficult to build.

3. Very little power is required from the noise source generator since the detector is quite sensitive.

On the negative side, you might expect some difficulty with this circuit for the following reasons:

1. The capacitance dial does not read out directly in ohms since this parameter is a function of the measurement frequency.

2. The resulting $R_p$ and $C_p$ values are parallel-circuit values, and series parameters are required in many practical applications.

3. A large amount of parallel capacitance is required to achieve balance with some modestly reactive impedances, particularly at low frequencies.

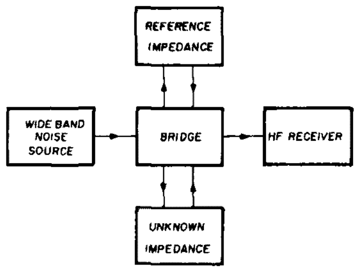



**fig. 1. The noise bridge includes five major elements, three of which are internal to the instrument. The heart of the instrument is the bridge section itself. This is excited by a broad spectrum noise source. The unknown impedance and a reference impedance form separate legs of the bridge section. The reference impedance is varied until it equals the unknown impedance. When this occurs, the bridge is nulled and the output of the high-frequency receiver goes to a minimum.**

Don't become overly discouraged over the negative aspects — we would like to add that none of these are of serious concern and, in fact, the first two reduce to simple calculations with a hand-held calculator. The third may be handled with the aid of a range extension assembly which we discuss in a later section. The required mathematical relationships will be covered in a section devoted to practical applications.

## initial units

Our exposure and efforts on the practical noise bridge circuit had an innocent beginning when we built the noise bridge described by YA1GJM with the modification suggested by K2BT.[1,2] Initial experiences were

*Rms stands for root-mean-square which is a measure of dispersion. Mathematically it is equal to the square-root of the average or mean value of the squares for a series or data points. For random errors, usually about two-thirds of the errors will be smaller than the rms error.

delightful. Extremely deep and repeatable nulls could be obtained while measuring complex impedances. Actual use of the device on antenna measurements and adjustment of a transmatch were not, however, nearly as fruitful. Adjustment of a transmatch to obtain a measured 50-ohm resistive impedance with a 21-MHz antenna system did not coincide with the adjustments to

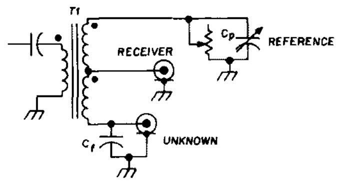



**fig. 2. The bridge section of previously recommended noise bridge designs included a trifilar wound transformer. This injects wide spectrum noise approximately equally into the two halves of the secondary. When the parallel combination of the unknown impedance, shunted by a capacitor, $C_f$, is equal to the parallel combination of $R_p$ and $C_p$, the bridge is balanced and the receiver detects a null condition.**

achieve minimum swr meter readings. Similarly, we found that the familiar Heath dummy load did not appear to be close to the 50-ohm design value at the higher frequencies. Somewhat perplexed at this point, we fell back to consider the next step.

K2BT mentioned evidence of calibration shifts when using the device at higher frequencies and this seemed to be related to our experience. We also began to suspect that our knowledge of the impedance of carbon resistors which we had been using to evaluate performance of the bridge was not satisfactory.

We decided to explore the intrinsic performance of the noise bridge in a more rigorous manner. Fortunately, we have access to a Boonton Radio RX-Meter, model 250A. This is a highly respected, laboratory quality instrument which also measures the R-C parallel equivalent representation of an rf impedance. Test circuits were constructed of various resistances, capacitances, and/or inductance combined in parallel or series. These were measured first on the Boonton and next on a second noise bridge unit which we built with more care to shorten leads in the bridge circuit. The results are summarized in table 1.

These data were mostly concentrated at the higher frequencies (21 and 28 MHz) where problems, if encountered, were expected to be most pronounced. Table 1 indicates some interesting consistencies as well as some interesting anomalies. The noise bridge nearly always indicated a more inductive (smaller) $C_p$ than the RX Meter. The only exception was test circuit 1. Here the error was not only capacitive but also extremely large. There was also a tendency for the noise bridge to indicate a higher than correct value for $R_p$, particularly for higher values of load resistance.

It is interesting to compare these results with the words of K2BT.[2] He thought his unit exhibited an "inductive rotation" of the $C_p$ calibration at the higher frequencies. That is a succinct way of summarizing our $C_p$ data in **table 1**. Conversely, however, K2BT did not notice $R_p$ errors at the higher frequencies. YA1GJM claimed that 2-watt composition resistors in the 10- to 150-ohm range exhibit "about ~14 pF inductance in parallel with their indicated resistance values."[1] Test circuits 1 and 4 are composition resistors, number 1 being a 15-ohm value, and number 4 a 150-ohm value.

We find it very interesting that our noise bridge also indicated roughly this value of shunt inductance for these resistors. Note, however, that the readings are *incorrect*! In general, our findings indicate that composition resistors above 50 ohms exhibit relatively small values of shunt reactance. In fact, these are handy devices for use in calibrating the noise bridge as we shall explain later.

We were basically dissatisfied with the performance of this latest noise bridge. We think it is fair to say that our unit was probably as good as those made by either YA1GJM or K2BT. Indeed, as pointed out above, there appear to be significant correlations in the way our unit performed with the descriptions provided by these authors. Therefore, it seemed probable to us that some systematic errors were inherent to all these units. This prompted us to vigorously attack the problem of accuracy improvement.

## improving accuracy

We could easily fill this entire issue of *ham radio* if we tried to relate, in any detail, all the avenues pursued in trying to improve noise bridge accuracy. A total of six different units were built with variations in each which we thought would be helpful. Many of our initial recipes for improving accuracy turned out to be blind alleys; others showed promise but did not provide the desired improvement. In the end we found four distinct design changes which definitely improve the performance of the instrument.

All of our improvements relate to the fundamental problem of achieving balance in the bridge section. Consider **fig. 2** which shows the previously recommended trifilar-wound transformer, T1. Clearly the desire is to balance both sides of the secondary circuit. Indeed, if the parallel combination of the unknown impedance and the fixed capacitor, $C_f$, is equal to the parallel combination of $R_p$ and $C_p$, the secondary should be balanced. However, this will be true only so long as there are no additional coupling paths which affect secondary circuit balance. Unfortunately, the toroidal transformer couples energy between the primary and the secondary circuits *both* inductively and capacitively. Thus, if the primary circuit is unbalanced as it is in the trifilar configuration, then you can expect this capacitive coupling to slightly unbalance the secondary circuit as well.

To understand this point, imagine small fixed capacitors from the primary circuit to the secondary circuit.

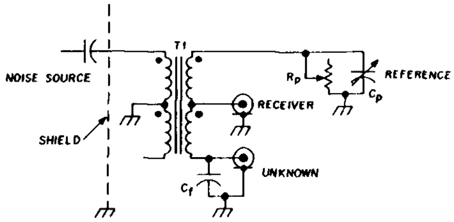



**fig. 3. The recommended improved bridge section contains a quadrifilar wound transformer. This winding technique significantly reduces unbalanced capacitive coupling between the primary and secondary circuits. An electrostatic shield between the bridge and the amplifier electronics is also recommended.**

Since the *dot* ends of each winding are physically near each other, then you expect the *reference* side of the secondary circuit to be most tightly coupled to the *ungrounded* side of the primary, whereas the *unknown* side would be more tightly coupled to the *grounded* part of the primary, circuit. It is not difficult to imagine rather significant unbalancing effects from this unbalanced capacitive coupling — particularly so at the higher frequencies.

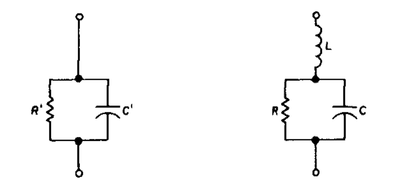
**Fig. 3** schematically shows the recommended transformer configuration. The toroid is wound in a quadrifilar configuration. The additional winding serves to balance the primary to secondary circuit capacitive coupling. Note that one end of this primary circuit is left floating. To achieve the most perfect balance, it should be tied back to ground through an impedance equal to the driving impedance of the amplifier stage. However, from a practical standpoint, our results indicate no measurable improvement in trying to simulate this amplifier impedance with various terminations. Hence, we recommend leaving the other end floating as shown in **fig. 3**.

Note also that a grounded electrostatic shield is installed between the amplifier and the toroid. This serves the same purpose as above, reducing stray capacitive coupling to the secondary circuit. These are two of the four improvements we made, and in retrospect it is clear that the quadrifilar winding is the more important improvement.

Ground loops can easily exist in rf circuits. Unfortunately, they are difficult to diagnose and their effects on circuit performance are sometimes difficult to predict. We had observed strange behavior in several of our early noise bridge models which we now ascribe to this phenomenon. The mechanism is relatively simple. The chassis is necessarily part of the secondary circuit. It is also possible to have currents from the primary circuit return through the chassis; however, when this occurs, the primary and secondary circuits are again coupled in a way which may lead to unbalanced behavior. Rather than pre-empt the section which carefully details how to build our improved noise bridge, we will conclude discussion of this third recommended improvement by simply saying it is necessary to carefully ground the amplifier and bridge circuit components to achieve the best accuracy.

After making all three of the above noise bridge improvements, we were still noting some residual high-frequency unbalance. Our results, measuring carbon resistors in the 150-200 ohm range with the Boonton 250A, indicate that these resistors have very small equivalent parallel capacitance — usually 1 or 2 negative picofarads. Hence, as we shall see in the section on checkout and calibration, they are convenient standards for estimating noise bridge accuracy. At 3.5 MHz, measuring a 150-ohm resistor, the modified noise bridge would correctly indicate very small parallel capacitance, but at 28 MHz it could indicate nearly −10 pF and, furthermore, the indicated resistance value was higher than it should be. We soon discovered that by reversing the secondary winding of the quadrifilar transformer, the effect could be made to reverse. Hence, the error at 28 MHz would now be in the capacitive direction and the indicated resistance would be too low. This clue was sufficient to point us in the right direction.

Consider fig. 4. Suppose we have a parallel R-C circuit to which we add a very small series inductance. We wish to find the parallel $R'C'$ circuit which, at some particular frequency, has the same impedance as the circuit with the added series inductance. The algebra you have to grind through to get expressions for $R'$ and $C'$ in terms of $R$, $L$, $C$, *and frequency* is a bit tedious. However, if you are willing to make the approximation that the




$$Z_0 = \frac{R'}{1+j2\pi fR'C'} \qquad Z_1 = j2\pi fL + \frac{R}{1+j2\pi fRC}$$

$$Z_0 \approx Z_1 \text{ IF } R' = \frac{R}{1+8\pi^2 f^2 LC} \text{ AND } C' = C(1+4\pi^2 f^2 LC) - L/R^2$$

**fig. 4. A small amount of series inductance added to a circuit results in a shift in the R-C parallel representation of the impedance of that circuit. The shift in impedance can be represented by another parallel R′ - C′ circuit. Numerical values for the shifted resistance and capacitance are found from the equations in the text. These are approximate in that they assume the reactance of the series inductor to be smaller in magnitude than the reactance of either R or C.**

reactance of the small inductor is much less than that of any of the other components, then a relatively simple result emerges:

$$R' \approx \frac{R}{1+8\pi^2 f^2 LC}$$

$$C' \approx C(1 + 4\pi^2 f^2 LC) - L/R^2$$

where $f$ is frequency.

Now to the point of this whole discussion. Suppose you wished to measure the impedance of a 100-ohm resistor with your noise bridge. On the unknown side you would have the 100-ohm resistor in parallel with the fixed capacitor. For the moment, assume the fixed

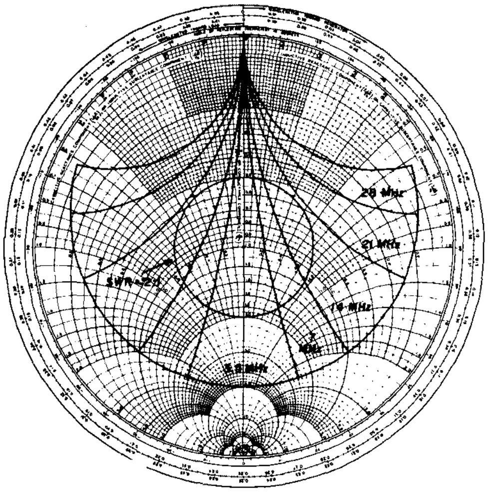



**fig. 5. The range of impedance which can be measured using the basic noise bridge configured with a 365 pF variable capacitor and a 250-ohm potentiometer is shown in this diagram. Since capacitor reactance varies with frequency, the range of the bridge increases with increasing frequency as shown. The center of the Smith chart is 50 ohms and other impedances are scaled relative to this value. A 2:1 swr circle is shown on the chart for comparison.**

capacitor to be 180 pF. Ideally, the bridge section would balance if you turned the pot and the variable capacitor on the reference side to these same values. Suppose, however, that one of these sides, either the reference or the unknown, had a small 10 nanohenry inductance in series with it. At 3.5 MHz, using the equations shown above, $R' = 99.8$ ohms and $C' = 179.2$ pF.

Hence, at this frequency, the very small series inductance causes a small but essentially immeasurable shift in bridge balance. However, at 28 MHz, using the same values as before, $R' = 90.0$ ohms and $C' = 189.0$ pF. Thus, the very small series inductance becomes a significant bridge unbalance at this frequency. If it exists on the known side, it causes the bridge to overestimate the resistance and to miss the correct capacitance on the inductive side. If it exists on the unknown side, the errors are reversed.

To put this in perspective, no. 28 (0.3mm) wire has an inductance of about 9 nanohenries/centimeter. In the process of winding a toroidal transformer and in wiring the bridge section it is not hard to imagine getting an extra centimeter of copper in one side of the bridge. Hence, our fourth and last suggested improvement is to achieve final bridge balance using a short piece of bare hook-up wire on the unknown side of the bridge to balance this effect. We will discuss this procedure in more detail in the section on checkout and calibration.

## range extension

In terms of parallel equivalent circuits, the range of the noise bridge using a 250-ohm pot for $R_p$, a 365-pF variable for $C_p$, and 180 pF for $C_f$ is roughly:

$$0 \leqslant R_p < 250 \text{ ohms}$$
$$-180 \text{ pF} \leqslant C_p < 180 \text{ pF}$$

This representation, however, is not very informative. Instead, let's look at this range as a contour plot on a Smith chart. W1DTY discusses the use of these charts in the November, 1970, issue of *ham radio*, and this discussion assumes that you are familiar with using them.[3]

Since the reactance of a fixed capacitor is frequency dependent, we can expect the range of the bridge to also be a function of frequency. This is shown in fig. 5. In this presentation the center of the chart is 50 ohms. Five contours are shown; each represents the range of the noise bridge at the indicated frequency.

Suppose you wished to measure the impedance of an antenna at 3.5 MHz. Suppose further that the antenna swr at this frequency is 2:1. Therefore, its impedance is somewhere on the 2:1 swr circle on the Smith chart. Note, however, from fig. 5 that only a very small portion of the possible impedances on a 2:1 swr circle are within the bridge's range at 3.5 MHz. K2BT noted this same problem, and recommended using a 365 pF variable instead of the 140 pF variable originally suggested by YA1GJM to increase the bridge's range. K2BT also suggested building the bridge so that fixed capacitance

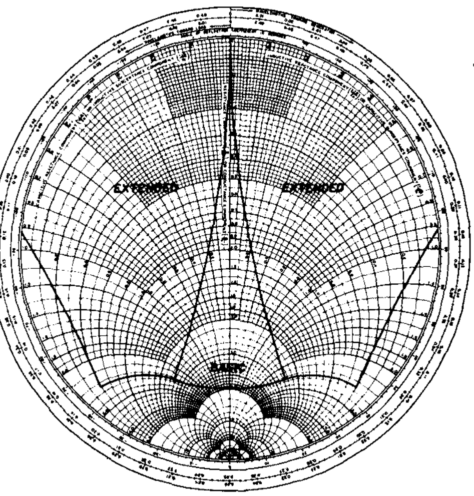



**fig. 6. The range of the basic noise bridge can be greatly extended by adding a 100-ohm resistor in series with the unknown impedance. The extended range of impedances which can be measured using this technique is shown on this Smith chart. The center of the chart is 50 ohms, and the extended range includes all points on and within a 5:1 swr circle at 3.5 MHz.**

could be added in parallel to either the reference or unknown sides of the bridge. We feel this is an excellent technique. Its only limitation is the required size of the fixed capacitance and possible construction difficulties. For instance, to measure any point on a 2:1 swr circle at 3.5 MHz requires the addition of up to 510 pF to the bridge. Extending the requirement to a 5:1 swr circle requires up to 2010 pF fixed capacitance.

An alternative which we feel is often more attractive is the addition of 100-ohm resistor in series with the unknown. In effect, this lowers the Q of the circuit to be measured and brings the resultant within the range of the bridge. The possible range extension at 3.5 MHz is shown graphically in **fig. 6**. This extended range includes all points on and within a 5:1 swr circle at 3.5 MHz and becomes even larger with increasing frequency.

A convenient range extender assembly will be described in the construction section; the mathematical formulas required to process the noise bridge readings with the 100-ohm range extender in place are included in the applications section.

## results

So far we have stated that accuracy improvements to the noise bridge can be made with certain design changes. We have also alluded to a technique for extending its range. The last three sections of this article will describe in detail just how to accomplish these results. However, before we proceed with these detailed descriptions, it's logical (and motivational) to demonstrate our claims.

**Table 1** summarized data taken on a noise bridge in accordance with earlier prescriptions. By way of comparison, **table 2** summarizes measurements made on these same terminations, but taken with two units built as described in this article. These are labeled, respectively, Unit 2 and Unit 3. Each of us built one of these, and other than the specific requirements imposed in the construction section, they are dissimilar. For purposes of comparison, the data from **table 1** (and Unit 1) are repeated.

Units 2 and 3 show a marked improvement in their correspondence to the Boonton measurements. The "inductive rotation" error evident in Unit 1 all but disappears in both these units. Further, the rather sizable error in measuring the $C_p$ of the first load has disappeared. The only significant errors which remained in the Unit 2 and 3 data are on the last test circuit where errors of 12 and 17 pF, respectively, were made. But these are one-half the corresponding error in Unit 1. Also, any substantial indication of a systematic error in resistance measurements has disappeared. The rms error pretty well summarizes the data. Units 2 and 3 outperformed Unit 1 by a factor of more than three to one!

Although the results shown in **table 2** pretty clearly indicate an improvement in measurement accuracy for both Units 2 and 3, we felt that additional data was desirable to more fully explore the accuracy potential of the improved units. Using a number of different building blocks, including discrete resistors, capacitors, and inductors and, in several cases, short coaxial lines with resistive termination, we built a number of additional test loads. Our intention was that these loads should include both inductive and capacitive reactance and both small and large values of resistance. They were synthesized to be roughly arranged around the 3:1-swr circle on a Smith chart with $Z_o = 50$ ohms. Some of

**table 1. A noise bridge unit built according to previously described articles as compared to a Boonton 250A RX meter.**

**MEASURED IMPEDANCE PARAMETERS**

| test circuit number | freq. MHz | Boonton $R_p$ | Boonton $C_p$ | noise bridge $R_p$ | noise bridge $C_p$ | error relative to Boonton $\Delta R_p$ | error relative to Boonton $\Delta C_p$ |
|---|---|---|---|---|---|---|---|
| 1 | 7.2 | 15.5 | −156 | 17 | −18 | 1.5 | 138 |
| 2 | 28.2 | 138 | −62 | 160 | −65 | 22 | −3 |
| 3 | 21.2 | 162 | −43 | 162 | −52 | 0 | −9 |
| 4 | 28.2 | 149 | −1 | 160 | −15 | 11 | −14 |
| 5 | 14.2 | 141 | 59 | 145 | 45 | 4 | −14 |
| 6 | 21.2 | 37.5 | 182 | 35 | 150 | −2.5 | −32 |

**table 2. Improved noise bridge units 2 and 3 as compared with unmodified unit 1 and the Boonton RX meter.**

**MEASURED IMPEDANCE PARAMETERS**

| test circuit number | freq. MHz | Boonton $R_p$ | Boonton $C_p$ | Unit 1 $R_p$ | Unit 1 $C_p$ | Unit 2 $R_p$ | Unit 2 $C_p$ | Unit 3 $R_p$ | Unit 3 $C_p$ |
|---|---|---|---|---|---|---|---|---|---|
| 1 | 7.2 | 15.5 | −156 | 17 | −18 | 14 | −160 | 16 | −160 |
| 2 | 28.2 | 138 | −62 | 160 | −65 | 142 | −55 | 140 | −56 |
| 3 | 21.2 | 162 | −43 | 162 | −52 | 161 | −41 | 160 | −40 |
| 4 | 28.2 | 149 | −1 | 160 | −15 | 150 | −4 | 147 | −1 |
| 5 | 14.2 | 141 | 59 | 145 | 45 | 142 | 60 | 136 | 58 |
| 6 | 21.2 | 37.5 | 182 | 35 | 150 | 35 | 165 | 34 | 170 |
| rms measurement error relative to the Boonton | | | | 10.2 | 58.5 | 2.1 | 7.8 | 2.9 | 5.9 |

**table 3. Improved noise bridge units 2 and 3 as compared with the Boonton RX Meter for a number of representative test circuits. All recorded data was converted to series impedance and therefore the entries have units of ohms.**

| test circuit number | freq. MHz | 100-ohm resistor used | measured impedance: Boonton | Unit 2 | Unit 3 |
|---|---|---|---|---|---|
| 1 | 7.2 | no | 15 + j2 | 14 + j1 | 16 + j2 |
| 2 | 14.2 | yes | 19 + j48 | 22 + j49 | 19 + j49 |
| 3 | 28.2 | no | 42 + j63 | 49 + j67 | 48 + j66 |
| 4 | 21.2 | no | 86 + j81 | 91 + j80 | 93 + j79 |
| 5 | 28.2 | no | 149 + j6 | 148 + j16 | 147 + j4 |
| 6 | 14.2 | no | 91 − j67 | 90 − j67 | 91 − j63 |
| 7 | 3.7 | yes | 51 − j45 | 52 − j44 | 51 − j44 |
| 8 | 21.2 | no | 21 − j19 | 22 − j17 | 21 − j17 |
| 9 | 3.5 | yes | 17 + j16 | 20 + j17 | 18 + j15 |
| 10 | 7.0 | yes | 23 + j34 | 28 + j35 | 27 + j32 |
| 11 | 14.0 | no | 120 + j61 | 122 + j58 | 123 + j54 |
| 12 | 21.0 | no | 55 − j59 | 54 − j61 | 54 − j58 |
| 13 | 28.0 | no | 21 − j16 | 21 − j17 | 19 − j13 |
| 14 | 3.5 | yes | 85 − j67 | 81 − j66 | 80 − j66 |
| 15 | 7.0 | yes | 39 − j51 | 38 − j53 | 38 − j49 |
| 16 | 14.0 | yes | 19 − j14 | 17 − j13 | 19 − j14 |
| 17 | 21.0 | yes | 20 + j14 | 22 + j12 | 20 + j16 |
| 18 | 28.0 | no | 43 + j47 | 48 + j49 | 46 + j52 |
| rms measurement error relative to the Boonton | | | | ΔR = 3.2 ohms<br>ΔX = j2.9 ohms | ΔR = 3.0 ohms<br>ΔX = j2.8 ohms |

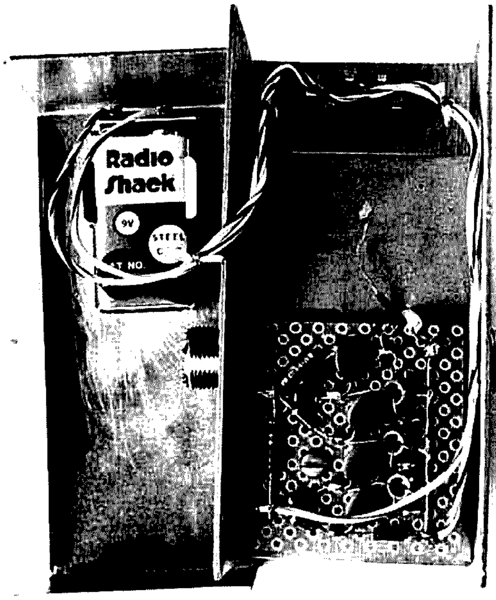



**fig. 7. Interior layout of the improved noise bridge is shown in this photograph. A copper electrostatic shield encloses the bridge section. The three leads from the bridge section can be seen at the upper right-hand corner of the perf board. One is a ground lead and the other two are leads from the primary of the toroid. One primary lead is connected to the amplifier output; the other is left unattached as explained in the text.**

these impedances are clearly outside the range of the noise bridge in its usual configuration. In these cases, the 100-ohm resistor technique briefly described above is required to obtain a measurement.

Table 3 shows the results of measuring these new test loads on Units 2 and 3. In all cases, the measurements were converted into the series circuit equivalent of the measured impedance. As before, the data are compared with those taken on the Boonton 250A RX-Meter. For convenience, these data were also converted to series equivalent.

There are several ways of describing the accuracy of these measurements. The rms error for each unit was about 3 ohms in both the resistive and reactive components. Considering the distribution of errors, one is 10 ohms, several are in the 5 to 7 ohm range, but *nearly half are 1 ohm or less*! While we shall purposely avoid making a global statement about the accuracy of these units, the data is encouraging well beyond our original hopes. Many high-quality instruments have accuracy expressions relating to "percent of full-scale reading;" our noise bridges appear to be in the "few" per cent category.

## constructon

Figs. 7 and 8 shows the details of construction we recommend. The 2¼ by 2¼ by 5-inch (5.7x5.7x12.7cm) box allows easy construction as well as a component arrangement which minimizes lead length. A conventional 365-pF air-dielectric variable can be made to fit in this same box, although it is recommended that the more compact Archer 272-1341 capacitor be used to improve access to the actual bridge circuit wiring during calibration. The amplifier circuit, fig. 9, was built on a small piece of perf board in the model pictured although two other units we built used a printed-circuit board. Both construction techniques resulted in identical performance and the circuit itself is immune to layout variations. The potentiometer should be a linear taper, carbon variety such as an Allen-Bradley, type J. Wirewound pots are unacceptable because of their high inductance.

Shielding is arranged so that the battery, electronics, and on/off switch are external to the critical bridge circuit components. Copper sheeting was formed so that one end is secured under the variable resistor and the sides are clamped between the flange of the chassis-mount coax connectors and the box. The electronics board was then mounted with a single screw and spacer on the major surface of the copper shield.

Short leads are important in the bridge circuit area with the exception of the leads going through a single hole in the shield near the output end of the electronics board. Of particular importance is the detail involved with grounding. The ground lug of the variable capacitor is used as the focus of a single-point grounding system; an insulated ground wire is routed from this point to the electronics along with both ungrounded primary leads from the toroidal transformer. Both ends of the primary

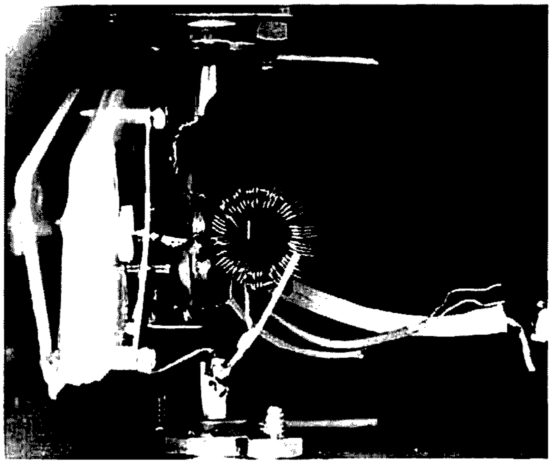


**fig. 8. This closeup photograph of the bridge section shows the location of major components. The small 365 pF variable is on the left while the 250-ohm pot is in the background behind the toroid. The SO-239 coaxial connectors are on either side of the box between the capacitor and resistor, providing a compact layout. The leads to the amplifier ground and from the toroid primary, as well as the hole through the copper shield to accomodate them, are in the lower right. The series inductance balancing device can be seen emanating from SO-239 connector on the bottom.**

windings will be required during the calibration sequence. Overall construction of the noise bridge is shown in **figs. 7** and **8**, and retention of this component layout is encouraged.

Construction of the 100-ohm range extender is quite simple with the use of a PL259-to-Motorola type pin plug (Archer 278-208). This unit fits conveniently into the back of a PL-259 plug (Archer 278-205). The adapter plug must be modified to allow for series connection of the 100-ohm, ½-watt carbon resistor. All but ¼ inch (6.5mm) of the Motorola plug is cut off, including the center pin. The 100-ohm resistor is soldered to the shortened inner pin and the assembly is then inserted into the PL-259 plug.

Several spots can be soldered to secure the two units together and the resistor lead projecting out of the PL-259 plug may then be soldered. The layout of the assembly is shown in **fig. 10**; the completed unit is pictured in **fig. 11**. It might be tempting to consider installing the 100-ohm range extender inside the noise bridge with a switch to more easily select its use. This was tried, with mixed results, so we recommend you use the external device if comparable accuracy is desired.

## checkout and calibration

Initial checkout of the noise bridge can be accomplished by connecting the unit to your receiver and leaving the unknown coax connector unterminated. There should be generous amounts of noise anywhere between 3.5 and 30 MHz! Although the output falls off somewhat at the higher frequencies, an S-9 signal should be expected on 10 meters. No null can be expected under these conditions. By varying the R and C dials, however, some variation in the S-meter reading is normal. Failure to observe these results requires checks of wiring and measurement of bias levels on the zener and transistor terminals. Access to an oscilloscope is helpful for signal tracing, but it was our experience that voltmeter measurements are quite satisfactory.

The next step in the checkout is to see if nulls can be achieved and that the bridge circuit is basically working. Connect the range extender assembly to the *unknown* coax connector and short the output of the 100-ohm resistor to ground through a physically short connection. By placing the receiver on 80 meters with a short agc time constant, if available, it should now be possible to find an extremely sharp null on the S-meter by adjusting both the $R_p$ and $C_p$ dials. Since there will be some interaction between the adjustments of the dials, the sequence must be repeated until the minimum S-meter reading is obtained. This null should be all but absolute, that is, near zero on the S-meter.

Since the unknown being tested is near 100 ohms with little reactance, the resistance dial, which is linear, should read approximately 40 per cent of scale from the zero position, and the capacitance dial would be expected to read near the center of its range. Inability to obtain these results indicates a problem in the bridge circuit and transformer connections should be checked.

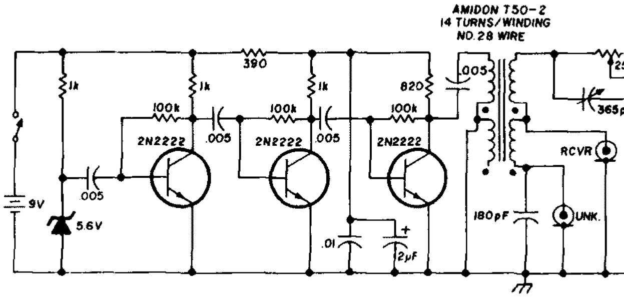



**fig. 9. Schematic diagram for the improved noise bridge unit. Construction details are discussed in the text.**

Offset of the capacitance dial from approximate center would indicate difficulty with the 180 pF fixed capacitor.

Assuming a successful checkout, you can now proceed with calibration. Preparation includes lightly steel-wooling the paint under the R and C dials to remove the gloss and permit the use of India ink or a fountain pen for scale markings. Later, a clear alcohol-based artist-type aerosol spray can be applied to protect the markings and make them more durable. Some clear sprays interact with ink, so a preliminary try is advised to avoid an upsetting experience after you have finished the calibration sequence.

With both the receiver and unknown disconnected, the resistance dial can be calibrated by attaching an accurate ohmmeter to either coax connector and ground. Surprisingly, dc resistance readings and accompanying marks on the resistance scale are very accurate from 80 through 10 meters. Tick marks every 10 ohms are a good balance between readability and scale appearance. Lacking an accurate ohmmeter, it might be advisable to locate several 1 per cent resistors in the 10 to 250 ohm range for reference.

Calibration of the capacitance dial is most easily accomplished with a capacitance meter. The variable capacitor must be temporarily disconnected from the rest of the bridge. Establishing the 180 pF dial setting as zero, the scale can then be marked in both directions to an "end-of-scale" reading of approximately 180 pF in either direction. Tick marks every 10 pF are desirable; the most clockwise 180 pF value should be marked with an L to indicate that inductive reactance is in that direction.

Lacking access to a capacitance meter, calibration of the C dial requires the following sequence: Connect the noise bridge to your receiver and select the 80-meter band. A 100-ohm, ½-watt carbon resistor and a selection of accurate capacitors in the 10 to 180 pF range should be at hand. Start with the 100-ohm resistor only and attach it to the unknown coax connector with short leads. Establish a null by observing the receiver S-meter and adjusting both the R and C dials. The resulting position of the C dial can then be marked as zero. By adding small values of capacitance in parallel with the

100-ohm resistor and rebalancing the bridge each time, the dial can be marked progressively until the 180 pF value is reached.

The next step is to temporarily remove the 180 pF fixed capacitor inside the noise bridge. Returning to the 100-ohm resistor measurement at 80 meters, capacitance should be added in parallel until the C dial reads zero as marked in the earlier step (approximately 180 pF will be required). Small values of capacitance should then be removed from the 100-ohm resistor until the scale has been marked as desired. Since the C scale is quite linear, some liberties may be taken in the form of estimating intermediate scale markings as long as accurate capacitors are used during the calibration process.

Following the scale marking sequence, all circuits can be returned to their original state in readiness for compensation adjustments which will assure accurate performance over the 3.5 to 30-MHz frequency. A 150-ohm, ½-watt carbon resistor should be placed on the unknown coax connector with short leads. The noise bridge may then be nulled on 3.5 MHz; both the resistance and capacitance dial should read fairly accurately at this point (150 ohms, 0 pF). Possibly the C dial might read slightly off zero at this time and the knob can be moved on the capacitor shaft to correct this situation. Moving your receiver to 30 MHz, again measure the 150-ohm resistor. If you get the same readings as you did at 3.5 MHz, the effort is over and you have a good instrument!

More than likely you will see a small shift at 30 MHz. If the $R_p$ dial reads higher, and the $C_p$ dial reads on the inductive side of zero, then your bridge needs additional series inductance on the unknown side. We recommend that you solder a short piece (about 1 inch or 25mm) of bare hookup wire to the SO-239 coaxial connector

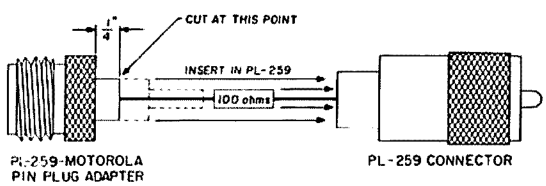



**fig. 10. Construction details for the 100-ohm range extender unit. A PL-259 to Motorola pin plus adapter is first prepared by cutting off all but ¼" (6.5mm) of the pin and shank. A ½ watt, 100-ohm resistor is then inserted in the pin of this adapter and soldered. The shortened shank and resistor are then inserted in the back of the PL-259 plug. The other resistor wire is soldered to the center pin of the PL-259 connector. Soldering the exterior surfaces together completes the assembly.**

marked *unknown*. The fixed $C_f$, 180 pF capacitor should remain soldered at this connector with a short lead to ground. The wire from the transformer secondary should be removed from the SO-239 connector and fastened instead to the 1 inch (25mm) piece of wire. Connect it to the far end of this wire and rerun you balance test.

You should now find that your error at 30 MHz has reversed (instead of $R_p$ reading too high, it will now be too low). In addition, the $C_p$ reading should be on the capacitive side of zero. If this is the case, move the wire from the transformer a little closer to the SO-239

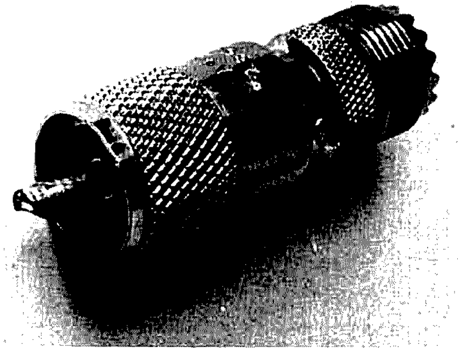


**fig. 11. The completed range extension assembly.**

connector along the 1 inch (25mm) wire and rerun your test. You should find a point where the $R_p$ and the $C_p$ dials read very nearly the same at 30 MHz as they did at 3.5 MHz. When they do, your work is complete.

Suppose, instead of $R_p$ too high and $C_p$ inductive at 30 MHz, your first measurement (without the 1 inch or 25mm wire) indicates the opposite. This means that your bridge has too much inductance on the unknown side. In this case, instead of trying to add wire to the known side it is easier to reverse the transformer primary windings to shift the error in the other direction. Then you can proceed to finish the calibration as discussed above.

Suppose you find a high-frequency shift that is different from those discussed above. Perhaps $R_p$ increases and $C_p$ also increases (goes capacitive). If this is the case, it indicates your unbalance is more than just a problem of unbalanced series inductance and you should check your layout, particularly the winding and placement of your transformer. Try and make this layout as clean as possible. Make sure your shielding is effective and that a chassis ground loop does not exist. If all this is done properly, the residual unbalance should be such that it can be compensated with the series inductance technique.

## applications

There are a number of applications in the shack which are ideally suited for the noise bridge. The first and most obvious is to use it to measure antenna impedance. Since the measurements will be made with the noise bridge located near the receiver, you will want to transform the values measured to those that apply to the antenna's feedpoint. Again, the Smith chart is

recommended and W1DTY's article or the latest issue of the ARRL *Antenna Handbook* discusses the technique in detail.[3,4] An example of the results that might be obtained are shown in **table 4** and **fig. 12**. These data are from an actual 80-meter inverted-vee antenna with isolation traps resonant on 40 meters. It is fed with 60 feet (18.3m) of RG-8/U coaxial cable.

The actual mathematical steps involved in noise bridge impedance measurements are quite easy and relatively fast once the pattern is established. First, record the bridge readings; then compute the reactance of the parallel capacitor. The parallel circuit elements are then converted to series equivalent elements. These transformations are performed using the following equations:

$$X_p = \frac{-159{,}000}{fC_p}$$

where $f$ = frequency in MHz
$C_p$ = capacitance in pF

For the 4.0-MHz data point,

$$X_p = \frac{-159{,}000}{4.0(-84)} = 473 \text{ ohms}$$

Converting to series equivalent values required use of the following equations:

$$R_s = R_p \frac{X_p^2}{R_p^2 + X_p^2} \qquad X_s = X_p \frac{R_p^2}{R_p^2 + X_p^2}$$

Substituting 4.0 MHz values of $R_p$ and $X_p$,

$$R_s = 117 \frac{(473)^2}{(117)^2 + (473)^2} = 110 \text{ ohms}$$

$$X_s = 473 \frac{(117)^2}{(117)^2 + (473)^2} = 27 \text{ ohms}$$

If the range extender was used, then actual values of the 100-ohm resistor must be subtracted from $R_s$ before it is entered in the corrected series impedance column.*

At this point, the equivalent series representation may be plotted on a Smith chart, making allowance for the rotation required due to the electrical length of the transmission line used. The electrical length is easily established by using the velocity factor ($v$) for the coax in use (66 per cent for RG-8/U) and the following relationship

$$\lambda = \frac{\ell f}{984v} \text{ wavelengths}$$

where $\ell$ = physical length of coax in feet
$f$ = frequency in MHz

For 4.0 MHz

$$\lambda = \frac{(60)(4.0)}{984(.66)} = 0.370 \text{ wavelength}$$

The equivalent series impedance components must, of course, be normalized to the $Z_o$ value (division by 50 in this case) prior to plotting on the Smith chart. Construction details for the 4.0 MHz data point are shown in fig. **12**. Although the influence of feedline loss could have been included, the degradation is small on 80 meters and, therefore, was ignored. The final results as presented by **fig. 12** allow considerable insight into the workings of this antenna:

1. Resonant frequency is 3.78 MHz.
2. Feedpoint impedance at resonance is 37 ohms.
3. Bandwidth with swr ⩽ 2:1 is 100 kHz.
4. Bandwidth with swr ⩽ 3:1 is 200 kHz.

With an antenna such as this one, you should not expect to load your transmitter at the band edge with-

**table 4. The noise bridge measured impedance of an actual 80-meter inverted vee antenna. Starting with the recorded values of $R_p$ and $C_p$, the data are converted to equivalent series impedance, and finally "rotated" through the 60-foot (18.3m) length of RG-8/U used to feed the antenna. The final column tabulates the measured impedance of the antenna at its feedpoint.**

| frequency MHz | noise bridge recorded data $R_p$ | $C_p$ | calculated capacitor reactance $X_p$ | series impedance $R_s$ | $X_s$ | 100-ohm resistor used | corrected series impedance $R_s$ | $X_s$ | transmission line length | antenna impedance $R_s$ | $X_s$ |
|---|---|---|---|---|---|---|---|---|---|---|---|
| 3.50 | 149 | −163 | 279 | 116 | 62 | yes | 16 | 62 | 0.323λ | 30 | −95 |
| 3.55 | 164 | −163 | 275 | 121 | 72 | yes | 21 | 72 | 0.328λ | 27 | −85 |
| 3.60 | 202 | −141 | 313 | 143 | 92 | yes | 43 | 92 | 0.333λ | 25 | −65 |
| 3.65 | 240 | −98 | 445 | 186 | 100 | yes | 86 | 100 | 0.337λ | 24 | −50 |
| 3.70 | 129 | −66 | 651 | 124 | 25 | no | 124 | 25 | 0.342λ | 26 | −31 |
| 3.75 | 78 | 152 | −279 | 72 | −20 | no | 72 | −20 | 0.346λ | 33 | −9 |
| 3.80 | 144 | 20 | −2092 | 143 | −10 | yes | 43 | −10 | 0.351λ | 41 | 9 |
| 3.85 | 121 | −9 | 4589 | 121 | 3 | yes | 21 | 3 | 0.356λ | 48 | 45 |
| 3.90 | 118 | −39 | 1045 | 117 | 13 | yes | 17 | 13 | 0.360λ | 65 | 70 |
| 3.95 | 116 | −65 | 619 | 112 | 21 | yes | 12 | 21 | 0.365λ | 75 | 105 |
| 4.00 | 117 | −84 | 473 | 110 | 27 | yes | 10 | 27 | 0.370λ | 95 | 140 |

*You should measure the actual value of your 100-ohm resistor by shorting the output end of the range extender and nulling your noise bridge.

out causing some sparks in your final tank circuit! A transmatch must be used to make the transmitter happy. Then there is the problem of properly tuning the transmatch at the frequency of interest. A good way to do this is to place the noise bridge at the transmatch input terminal and set the dials to $R_p$ = 50 ohms and $C_p$ = 0. Then by turning the transmatch dials, look for a null.

If a null can be found, the resulting swr at the transmitter will be very close to 1:1. And the best part is that you've done this without radiating any power and causing interference. It's a good idea to log these results for future reference. Be cautious, however, with this test setup, because it's possible, by hitting the wrong switches, to apply power to the antenna through the noise bridge. If you do this, even for an instant, you'll be in the market for a new 250-ohm pot and perhaps a new transistor in the amplifier circuit.

The particular antenna discussed earlier did not have a balun installed at the time these measurements were taken. In trying to find a suitable core to wind a balun, we discovered another application for the noise bridge. A core of appropriate dimensions but uncertain ancestry was available. In practice the resulting balun transformer didn't work. To discover why, a 50-ohm carbon resistor was soldered across the secondary of the balum and the input impedance measured with the noise bridge. With a good 1:1 balun we would expect to see an $R_p$ of approximately 50 ohms in parallel with a negative $C_p$ representing the reactance of the primary of the transformer. If this parallel reactance is large with respect to 50 ohms, the transformer works as it should. In this case we found a very high impedance primary circuit, beyond the range of the bridge. This indicated that the core material was not designed for the frequency range desired so a different core had to be obtained.

Another pet application involves power meters. We have built several of these and have wondered about their accuracy. If you have both a 50-ohm dummy load and a transmatch, you can use the noise bridge to synthesize arbitrary transmitter loads of interest. Unless you spent a lot of money for your dummy load, it probably is not exactly 50 ohms at any frequency. If you used it to set the null on your power meter, that null will not be the best that can be obtained. By using the transmatch/dummy load combination, you can synthesize an accurate 50-ohm impedance.

In addition, do you have confidence that when your power meter indicates a 2:1 swr that this is in fact the case? With the above hardware, you can synthesize an rf impedance of either 100 ohms or 25 ohms. Both should yield an swr of 2:1 on a 50-ohm coaxial cable. You might try this to see if your power meter indicates the correct values.

The last application we shall discuss effectively demonstrates the extreme range capability of the instrument when using the 100-ohm resistor modification. It is often desired to know the inductance of an rf coil. With air-core coils having uniform dimensions, it is possible to accurately compute inductance using

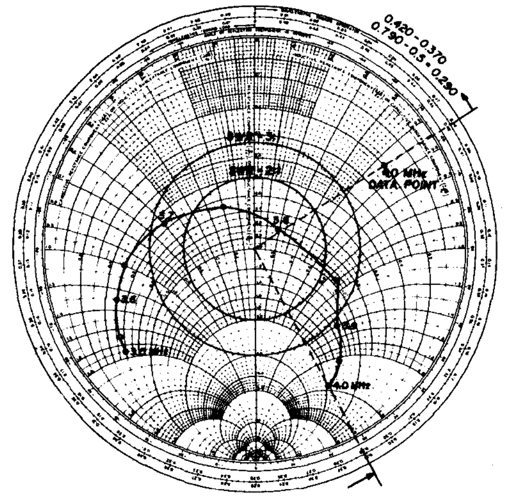


**fig. 12. Smith chart plot of the noise bridge measured impedance of an actual 80-meter inverted vee antenna. The construction of the 4.0 MHz data point is shown for illustration. The measured impedance at 4.0 MHz must be rotated 0.370 wavelength "toward the load" to account for the 60′ (18.3m) RG-8/U transmission line. 2:1 and 3:1 swr circles illustrate the bandwidth of the antenna. The bandwidth is relatively narrow since the antenna is physically short (about 100′ or 30m long) and relies on the inductive loading effect of 40-meter traps to achieve resonance on 80 meters.**

standard formulas. These formulas are much less precise and are more difficult to apply to inductors wound on a core such as a toroid. However, using the rf noise bridge, it is very easy to directly measure inductance in the "few" microhenry range.

By nulling the noise bridge with the 100-ohm resistor in the circuit and the unknown coil attached, observe the $R_p$ and $C_p$ readings. If a reading is not possible, try a lower frequency. The $C_p$ reading should be negative, of course, indicating the inductive reactance of the load. The inductance of the unknown coil in microhenries is calculated by

$$L = \frac{-R_p C_p}{10{,}000} \ \mu H$$

We leave it as an exercise to mathematically inclined and interested readers to derive this equation. Most, I'm sure, would prefer to prove it by demonstration with a real rf coil. We earnestly invite you to build the rf noise bridge as described here so you can perform the demonstration yourselves!

## references


1. G. Pappot, YA1GJM, "Noise Bridge for Impedance Measurement," *ham radio*, January, 1973, page 62.
2. Forrest Geherke, K2BT, "Impedance Bridges," (letter), *ham radio*, March, 1975, page 60.
3. James Fisk, W1DTY, "How To Use the Smith Chart," *ham radio*, November, 1970, page 16.
4. *The ARRL Antenna Book*, 13th edition, ARRL, Newington, Connecticut, 1974, page 76.


**ham radio**

# milliwatt portable counter

## A low-power frequency counter using RCA COS/MOS ICs that operates to 4 MHz and above

How would you like to have a truly portable milli-powered frequency counter that is usable to over 4 MHz? The RCA line of COS/MOS digital logic integrated circuits makes it possible to build such a counter with a total power consumption of 300 milliwatts (12 volts at about 25 milliamps including current to light a seven-digit display).

### features

With no circuit changes, such a counter can be run at any loosely regulated voltage between 4 and 15 volts with no loss of accuracy. It's possible to operate the counter from a common 9-volt transistor battery, a 12-volt automobile battery, or rechargable nickel-cadmium batteries. Obviously, line-power operation is possible by using a 6-volt transformer with a rectifier and filter.

The portability feature of this counter allows it to be used in hard-to-reach places in the same way that a hand-held vom is sometimes more useful than an ac-operated bench-type vtvm.

Although the frequency range of the counter is stated as "over 4 MHz" (to comply with published specifications on COS/MOS ICs), the model constructed actually works to 6.2 MHz. If a higher frequency range is desired, a single TTL divide-by-ten prescaler will increase the frequency to 50 MHz, which is more than adequate for most applications.

In designing this very low-power counter, it was decided to follow a nonconventional course of logic. This approach gives some insight into the use of a random-access memory, bus registers, and multiplexing.

A copy of the printed circuit board layout can be obtained by sending a self-addressed stamped envelope to *ham radio*, Greenville, New Hampshire, 03048

This last item, multiplexing, is of interest because it provides a great reduction in digit current. Each digit is illuminated only 10 to 20 percent of the time. In other words, each digit operates on a 10-percent duty cycle and is off more than on. To the eye, of course, it appears to be continuously on. The peak current is higher than the normal continuous current, but because of the human eye, the average current can be less for the same apparent brightness. In addition, only one digit is lit at any given time.* Also the use of multiplex reduces the number of wires between the circuit board and the digits themselves. This makes is easy to mount the circuit board near the frequency source to be measured while the digits are placed where they are most convenient to read.

### conventional counter

**Fig. 1** shows the block diagram for a conventional frequency counter. Each divide-by-ten IC has a BCD output, which is stored in a latch, decoded, and displayed. While the dividers are counting, the information in the latches keeps the digits lit with the previous data. Periodically the new data is fed into the latches and the digits show the new values. All digits are lit at all times.

In a multiplex system, the decoder outputs are tied physically to a common set of bus lines. At clocked intervals, each decoder becomes electrically connected to the buses and its data is fed to all the LEDs. However, at the same time, only one digit corresponding to the *on* data is allowed to light through its common emitter and an external transistor that is clocked in synchronism with the data. Thus multiplexing is ideal for remote readout of data collected elsewhere. Most decoder outputs are two-state; that is, either high or low. With several outputs tied together, a conflict of decisions is continuously occurring on the common bus lines. To separate the data, it is necessary to put a diode in each output line to isolate it except when its data is desired. For seven digits, each with seven segments, 49 diodes are needed.

### improved system

To overcome this disadvantage, while being able to reduce the number of wires by 65 percent, a different system was developed, still using RCA COS/MOS devices for low-power consumption. **Fig. 2** is a block diagram of this system.

The circuit difference lies in the manner in which data is combined, multiplexed into memory, then separated into two multiplexed outputs. One output is for the 3 least-significant figures and the other for the 4

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*In this circuit, two digits are lit at once as explained later.

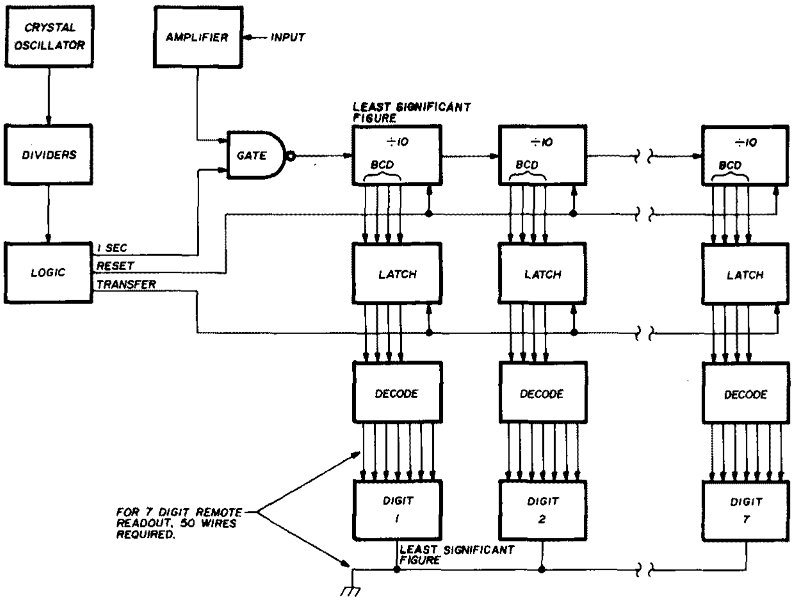



**fig. 1. Block diagram of a conventional counter. Fifty wires are required between decoders and digital readouts.**

most-significant figures. Use **fig. 2** or **3** to follow the explanation.

A 10-kHz crystal oscillator is used as the standard clock. The dividers and logic gates, U14 through U19 and half of U5, are used in conventional circuitry to provide the one-second count time as well as reset and write pulses. Note, however, that U15, an RCA CD4017A, has parallel as well as serial outputs. At any

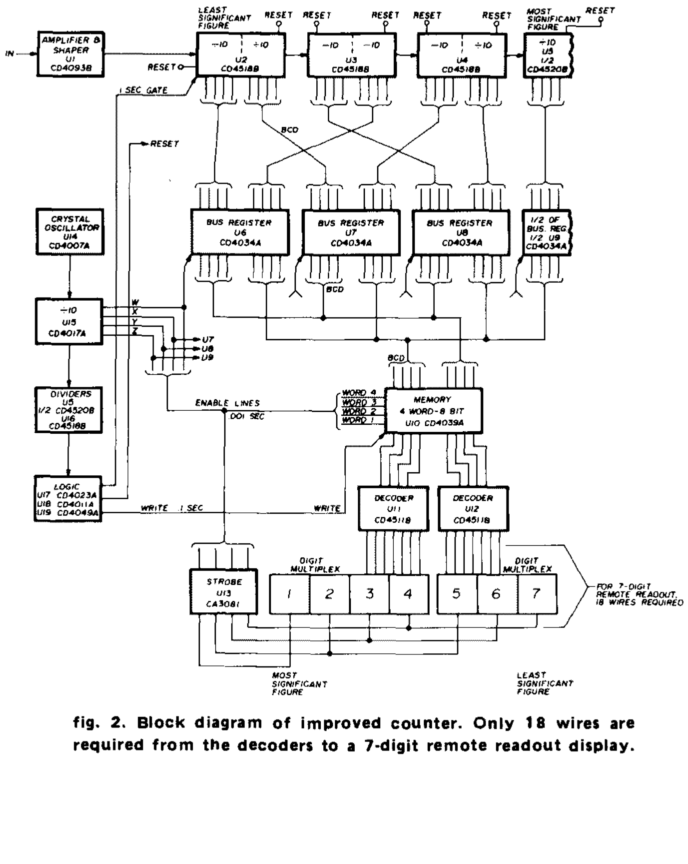



**fig. 2. Block diagram of improved counter. Only 18 wires are required from the decoders to a 7-digit remote readout display.**

time one of the 10 parallel output pins will be high. Each input pulse will move the high output in continuous sequence from zero to nine and back to zero. Each of these output pins will go high for every tenth pulse, but only one will be high at any given time. These pulses are the source of the clock timing for the bus registers, memory location and digit *on* time.

Assume the counters are counting. The changing BCD data in each counter also appears at the input to their respective bus registers, which are inactive until they are turned on. Each bus register has its own *enable* line tied to a different parallel output line of the CD4017A. One at a time, each bus register is activated, and the counter data is fed to the 8-bus multiplexer.

The memory, U10, an RCA CD4039A, has storage capacity for 4 words of 8 bits each. The memory sees this continuous changing flood of data on its input lines, but ignores it until it receives a *write* pulse from the control logic. This pulse occurs only after the counters have finished their one-second count and are in standby with the final count data. The location in memory for each data set is then chosen by a pulse that selects the word-enable lines, one at a time in synchronism with the data being fed to the bus lines. For one word, this data is the same pulse that enables bus-register one, U6. Therefore, word one becomes memory storage for the data from register one. The enable pulse then moves to bus register two, U7, and word two is enabled. The same action occurs with words three and four. This circuit has now multiplexed the counter data into memory.

Now let's notice the choice of data being fed to each memory word. The counter outputs are BCD and are therefore 4 bits long. Since each bus register can handle 8 bits and each word is also 8 bits long, the data from two counters can be fed simultaneously to each memory word.

The pairs chosen are not consecutive but are a

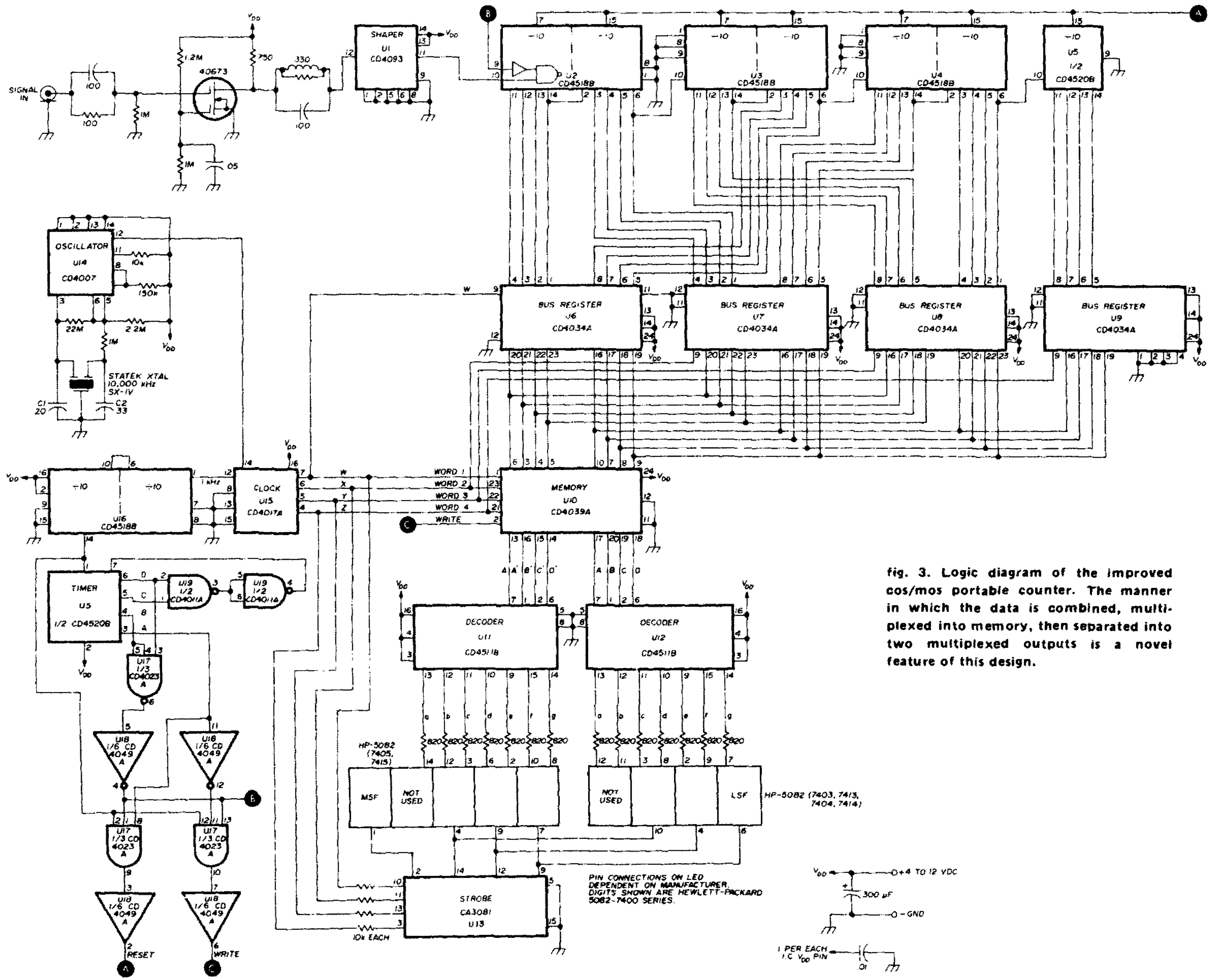



**fig. 3. Logic diagram of the improved cos/mos portable counter. The manner in which the data is combined, multiplexed into memory, then separated into two multiplexed outputs is a novel feature of this design.**

combination of digit one with four, two with five, and three with six; digit seven operates alone. The reason for this becomes evident when we see how the data is read from memory and separated.

Because the memory words each contain 8 bits (two BCD sets), two 4-bit decoders are necessary. One will receive the 3 least-significant figures and the other the 4 most-significant figures. The data is fed from memory by the same pulse that chooses the word to receive data. However, the reading ability is continuous while writing can occur only when the write line enables it. Thus the latest final-count data in memory is continuously being circulated to the output lines: first word one, then two, and so on. At any instant only one word appears at the output. As each word appears it is decoded by the two RCA CD4511 ICs U11 and U12. The first four bits go to one decoder; the last four to the other.

In a multiplex readout, the a segment of each digit is tied to a common line, each b to another line, and so on. The choice of the digit that is lit is made through the cathode of each digit. Thus digits 1, 2, and 3 have common segment lines. Digits 4, 5, 6, and 7 have a separate set of common segment lines. One of the decoders just mentioned goes to one multiplex set and the other to the second multiplex set. This is possible because the data being handled is from dividers one and four, U2 and U3. Had the data been from consecutive dividers (say one and two), the decoder outputs would have to feed data to the same line at the same time. As shown earlier this is not possible.

After word one is read and transferred to the digits, word two is read and data two and five are sent to the digits. Words three and four then follow in sequence. Since all three (or four) digits on a common set of lines receive the same data that was meant for only one digit, it is necessary to strobe each digit in time with its proper data. This means that the clock pulse that allows word one to be read must also turn on the correct pair of digits corresponding to this data. This is done by transistor, U13, an RCA-CA3081 connected to the cathode lines of each pair. The transistor is turned on by the same clock pulse that allows word one to be read.

The strobe rate is not critical as long as it's faster than 50 pulses per second to prevent flicker. In the circuit shown, it is strobing at 1 kHz because that was a convenient divider from which to draw the pulses U15.

Since any digit is on only one-tenth of the time, the peak current sent to each segment is higher than it would be for continuous operation, but the average current is actually less than it would be for continuous operation.

## construction

Construction should present no problems. The PC board is shown in fig. 4. A top view is shown in fig. 5. Because of the number of lines on the circuit board, the use of a single-sided board requires several cross-over wires. The more ambitious might try a two-sided board. Layout is not critical. Bypass capacitors, 0.01 $\mu$F, are used at the $V_{DD}$ pin of each IC where possible. These,

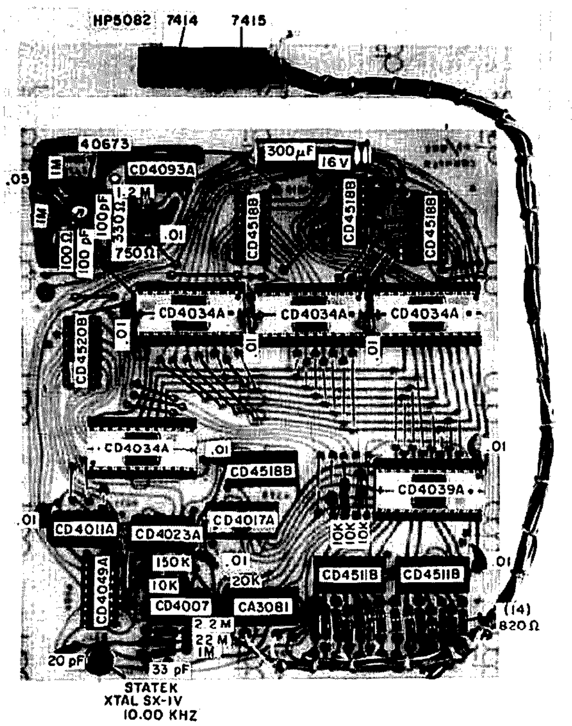



fig. 4. Top view of the layout of the portable counter.

together with the large power-supply filter capacitor, eliminate any oscillation problems. For those unable to make the board, the circuit can be hand wired.

There are absolutely no adjustments to be made for complete operation. Accuracy will depend entirely on the accuracy of the crystal clock. Use of a 0.01 percent accuracy crystal will give 0.01 percent accuracy ±1 digit. The clock frequency may be trimmed slightly at C1 or C2 (fig. 3).

## applications

The use of the counter is limitless. Complete portability with battery power allows its use in the home or field. It is excellent for setting teletype mark and space tones or fm repeater tones. Signal-generator frequencies as high as 4.5 MHz for TV alignment can be set more accurately. Another suggested use is musical instrument tuneup with a microphone pickup. The counter should be excellent for the piano or organ, especially electronic organs where direct pickup from the dividers is possible. With external circuits, it can be used in the car as the readout for a tachometer or speedometer. Surely you have already found a need for it.


## acknowledgement

I would like to acknowledge the help received from RCA COS/MOS Applications Engineering and especially from Stanley Niemiec in choosing the ICs used in the circuit.


**ham radio**

# interstage 50-ohm terminator for vhf converters

Many mixers and preamplifiers require a 50-ohm load — here's a circuit for providing a wideband resistive termination with minimum insertion loss

By H. Paul Shuch, WA6UAM, Microcomm, 14908 Sandy Lane, San Jose, California 95124

**When cascading modules** for receiving systems, it is often necessary to make sure that a particular stage is presented with a reasonably precise 50-ohm termination, over a relatively broad range of frequencies. One such requirement involves terminating a low-noise preamplifier which, although unconditionally stable at the operating frequency, is potentially unstable out-of-band. The highly reactive termination presented by a bandpass filter, operated off resonance, may cause the amplifier to oscillate at some undefined frequency, significantly degrading system noise figure and intermodulation performance. Another problem involves the image-frequency termination of double-balanced mixers. It has been shown that, to maximize a mixer's dynamic range, the i-f port must be properly matched – not only at the signal frequency, but also at any multiple-response frequencies appearing at the i-f.[1]

One method for obtaining a broadband interstage impedance match is based on the use of resistive attenuators between various stages.[2] Unfortunately, this approach introduces additional system losses which tend to degrade overall sensitivity. Another solution uses *interstage duplexers* which shunt the undesired frequency into a 50-ohm load.[3] This practice, however, is applicable only when the frequency of the undesired response is known and is well removed from the signal frequency. The circuit in **fig. 1** overcomes these shortcomings: It appears virtually lossless at the signal frequency and provides a wideband 50-ohm termination to any other frequency components which are present (limited only by the reactive nature of the load resistors at microwave frequencies). Additionally, this network provides the desired degree of interstage selectivity, as a function of the component values chosen.

The circuit of **fig. 1** is by no means original; it was brought to my attention by Gary Frey, W6KJD, who first encountered it in a commercial receiver design. Gary and I have both used the circuit extensively in vhf and uhf transmit and receive converters with considerable success.

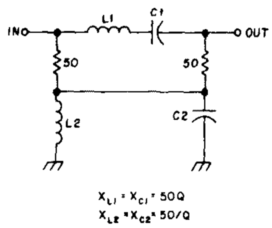



**fig. 1. Vhf terminator which provides a wideband 50-ohm termination with minimum insertion loss. Component values are based on desired circuit $Q$, as discussed in the text. Equivalent circuits at resonance, and above and below resonance, are shown in fig. 2.**

## circuit operation

In the circuit of **fig. 1** capacitor C1 and inductor L1 form a series-resonant circuit at the operating frequency, while C2 and L2 are parallel resonant. At resonance the impedance of L1-C1 is at a minimum, the impedance of L2-C2 is maximum, and the signal path from input to output appears as a short circuit across the two 50-ohm

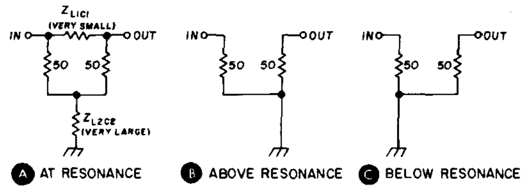



**fig. 2. Equivalent circuit of the 50-ohm vhf terminator at resonance (A), above resonance (B), and below resonance (C).**

resistors, as shown in **fig. 2A**. Thus the insertion loss of the network at resonance is minimal (due primarily to component losses in the resonant circuits).

At frequencies far above resonance, capacitors C1 and C2 appear as short circuits, and inductors L1 and L2 appear open. Thus the circuit is equivalent to that shown in **fig. 2B**, with input and output isolated from one another, and each port terminated in a 50-ohm load.

Well below the resonant frequency, both capacitors appear open, the two inductors may be thought of as short circuits, and the equivalent circuit of **fig. 2C** applies. Again, maximum isolation exists between the two ports, and each is terminated in 50 ohms.

A less clearly defined condition exists at frequencies slightly removed from resonance. Isolation is incomplete and the transfer coefficient is a function of circuit Q. Thus the selectivity characteristics of a single-pole band-pass filter are achieved. However, non-propagated signal components are not reflected, as would be the case with a simple resonant circuit. Rather, they are absorbed by the 50-ohm loads, giving the interstage network its wide-band terminating properties. Since reflected waves are not evident from either port, bilateral out-of-band isolation has been achieved.

## determining circuit Q

Assuming minimum dissipative losses in the reactive components, circuit Q is primarily a function of the ratio of the reactances at resonance to the terminating impedance (50-ohms in this case). Selecting a desired circuit Q, component reactances at resonance are found *from:*

$$X_{L1} = X_{C1} = 50Q$$

$$X_{L2} = X_{C2} = 50/Q$$

Ideally, any desired circuit Q could be selected, and component values derived. Practical considerations, however, restrict practical values of Q to 10 or less. Higher Q is possible if passband insertion loss is not a significant consideration, but this usually requires that variable capacitors be used to set the network to resonance at the desired frequency. With lower values of Q, fixed components of standard values may be used with minimum circuit degradation.

The required Q is a function of the amount of out-of-band isolation which is desired, as well as the frequency separation between the signal and spurious components. It is useful to relate isolation requirements to ripple bandwidth, which is defined as center frequency divided by Q. As a rule of thumb, isolation is 10 dB for frequency components separated *from* resonance by ±3BW and 20 dB of isolation is achieved at the center frequency ±10BW.

In receiving converters, when terminating the i-f port of a balanced mixer, the rf feedthrough, LO feedthrough, and image frequency components may be separated from the i-f signal frequency by an order of magnitude or more. In such cases a Q of one may be entirely adequate to effectively isolate all spurious components. (Incidentally, a Q of unity is the only case for which C1 = C2 *and* L1 = L2).

Improperly terminated uhf preamplifiers, on the other hand, often tend to oscillate in the vhf spectrum (a common occurrence with Microcomm's RA-70, 432-MHz preamplifier, for example). Therefore, the terminator following *a* preamplifier should exhibit relatively high Q so it will provide adequate isolation at the frequency of potential instability, thus suppressing oscillation. An acceptable compromise seems to favor a Q of about 5. Insertion loss thus remains low (fractions of a dB), selectivity is moderate, and components have practical values and are non-critical.

**Table 1** lists actual component values for terminators operating at various i-f and rf frequencies of interest to radio amateurs, assuming a circuit Q of 5. At the lower frequencies the circuits may be built successfully by using disc capacitors and either miniature molded rf chokes or hand-wound toroidal inductors. In the uhf region, the use of chip capacitors and microstripline inductors seems more appropriate. Of course, as frequency is increased, lead lengths must be kept to a minimum.

**table 1. Interstage 50-ohm terminator component values (Q = 5) for various vhf and uhf amateur bands.**

| | frequency (MHz) | | | | | | |
|---|---|---|---|---|---|---|---|
| | **10.7** | **28** | **50** | **144** | **222** | **432** | **1296** |
| L1 (nH) | 3720 | 1420 | 796 | 276 | 179 | 92 | 30.7 |
| C1 (pF) | 59.5 | 22.7 | 12.7 | 4.4 | 2.9 | 1.5 | 0.5 |
| L2 (nH) | 149 | 56.8 | 31.8 | 11.1 | 7.2 | 3.7 | 1.2 |
| C2 (pF) | 1490 | 568 | 318 | 111 | 71.7 | 36.8 | 12.3 |

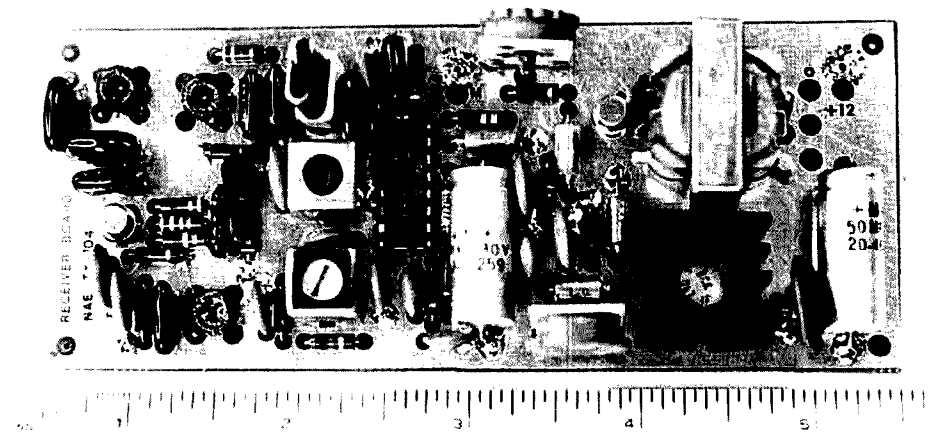
## acknowledgements

I wish to thank Gary Frey, W6KJD, for bringing this terminator circuit to my attention, explaining its operation to me, and calculating the component values presented in **table 1**. And I owe a special thanks to Stan Savage, W6ABN; his frustration in fighting oscillations in not one, but two Microcomm preamplifiers, convinced me of the importance of providing an effective, broad-band impedance match.

## references

1. Peter Will, "Reactive Loads — The Big Mixer Menace," *Microwaves*, April, 1971, page 38.
2. Edward L. Meade, Jr., K1AGB, "Using the Double-Balanced Mixer in VHF Converters," *QST*, March, 1975, page 12.
3. Doug DeMaw, W1CER, "His Eminence — The Receiver," *QST*, June, 1976, page 27.

**ham radio**



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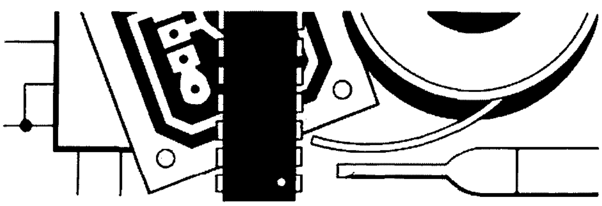


## fixed-frequency receiver for WWV

Calibration standards are essential to ensure on-frequency operation when working with rf circuits. One source of such standards is the National Bureau of Standards station WWV. In addition to frequency calibration data, WWV provides propagation forecasts, geophysical alerts, and storm information. All this information is available to anyone with a receiver capable of receiving WWV signals.

Some work I was doing recently with crystal oscillators required a reference frequency that WWV could provide. I didn't have a receiver capable of tuning any of the NBS station frequencies, so I decided to design and build one. Requirements were that the receiver have high sensitivity, portability, low-power consumption, and low cost. This article provides design and construction information for a receiver that meets these requirements and which, with a little innovation, can be readily adapted for other uses in the hf spectrum.

By Andrew M. Hudor, Jr., Physics Department, University of Arizona, Tucson, Arizona 85721

The RCA CA3088 IC[1] was chosen as the basis for the receiver design. The converter, i-f, detector, audio preamplifier, agc, and a tuning meter output are all contained on this single chip! The block diagram of the CA3088 is shown in fig. 1A. Using this IC resulted in a saving of size, cost, and design time.

The CA3088 is not without its disadvantages, however. A good front end design is critical to receiver performance, since any signal lost there can't be recovered later, and any noise introduced at that point is amplified with the signal in the following circuits. A separate mixer and local oscillator provide superior performance to the simple converter circuit in the CA3088. Fortunately, the converter and i-f input on the IC are not committed to each other. Fig. 1B shows the internal schematic of the CA3088; it can be seen that the converter transistor has all of its terminals accessible. In fact, it has dc bias applied to its base, and only the ac paths are required to make it a crystal-controlled local oscillator.

**Converter.** A modified Pierce oscillator was designed, using the converter transistor of the CA3088. A mixer circuit was taken from reference 2. This circuit uses a junction fet, which gives more gain than a diode mixer and has less noise than a bipolar transistor. Fewer components were required to bias the junction fet, which was the main reason for selecting it over a mosfet.

**Rf stage.** A dual-gate-protected mosfet was used for the rf stage. This circuit was also taken from reference 2. The mosfet has excellent gain and low-noise performance, and the dual gates allow one gate to be used for the incoming signal, while agc can be applied to the other. The mosfet also has the advantage of not normally requiring neutralization when used in small-signal rf amplifiers. This greatly simplified alignment while eliminating a costly and space-consuming trimmer capacitor.

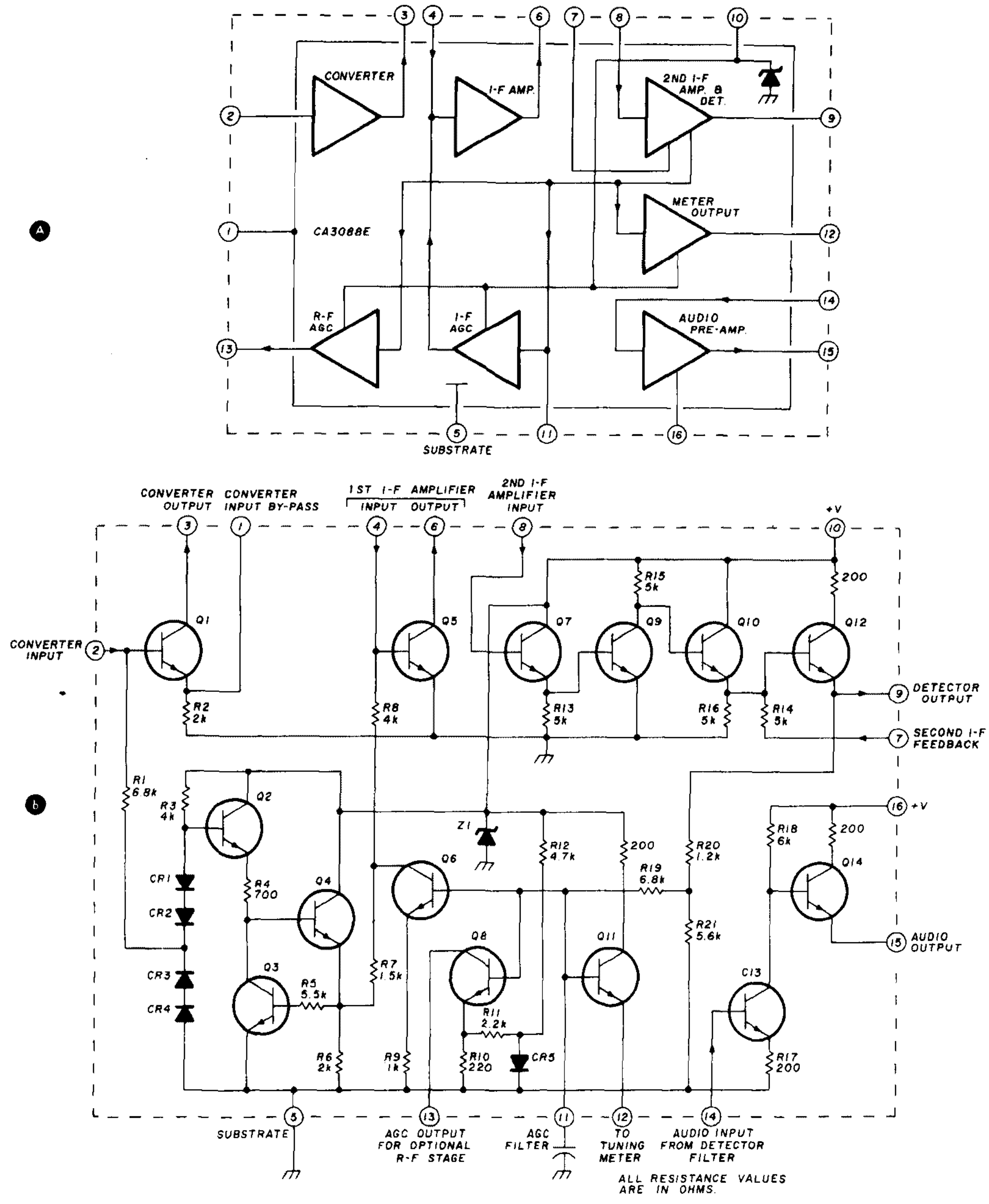


fig. 1. Block diagram of the RCA CA3088 receiver chip, A, and its internal schematic, B (from reference 1). In the WWV receiver design a separate local oscillator and mixer replace the converter circuit shown here.

**Coils.** The use of toroidal inductors also helped to reduce circuit complexity. The toroids have sufficient bandwidth so that adjustment is not necessary in the rf stage, mixer, or local oscillator. Although the very first version of the circuit used standard slug-tuned inductors, the toroids proved superior by making the circuit more stable. The area occupied by the inductor was reduced, adjustment was eliminated, and the toroids were actually lower in cost than the original coils.

**I-f, detector, and audio.** The i-f, detector, agc, and meter output were used as in reference 3. Transformers were used for the i-f stages instead of crystal filters since they were readily available and lower in cost.

The RCA CA3020 was selected for the audio-output stage, since I had a few of them in my parts stock. It turned out that using this IC yielded several advantages. The frequency response of the CA3020 can be shaped by selecting coupling and bypass capacitors. These components were selected to give an audio bandpass between 300 Hz and 3 kHz to reduce noise and increase signal intelligibility. The basic circuit for the audio output stage was taken from reference 4. The CA3020 had

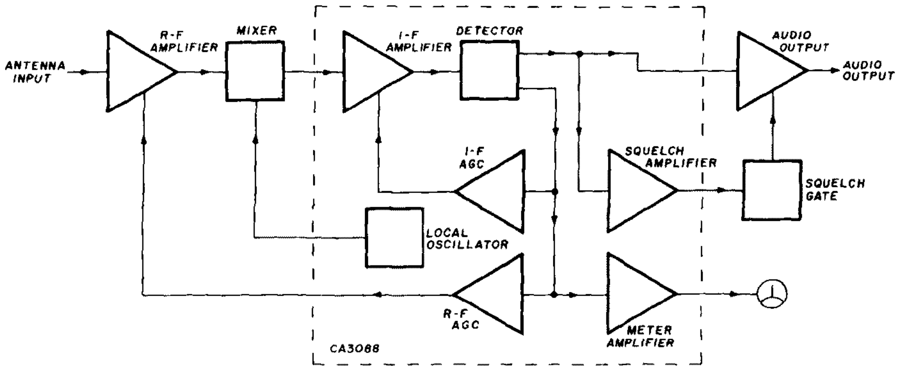



**fig. 2. WWV receiver block diagram. An RCA CA3020 was added as an audio-output stage. The af circuit has a shaped frequency response, which reduces noise and increases intelligibility.**

sufficient gain to be driven directly from the detector output of the CA3088. It also has squelch capability by application of a dc control signal to pin 11. The remaining audio preamplifier on the CA3088 was reconfigured, and with a few other components, a squelch circuit was added to the reciever.

The complete block diagram of the receiver is shown in fig. 2. The result of the design was a high-quality unit that has many other possibilities in addition to its orignial application as a fixed-frequency WWV receiver. Total parts cost was about $30.00.

## construction

The 10-MHz WWV signal was chosen for the prototype design, since this frequency has the best daytime signal strength in my area. The circuit, fig. 3, can be used in any part of the hf spectrum with only minor component changes. Table 1 gives component values for WWV, WWVH, and CHU frequencies. For local-oscillator frequencies below 20 MHz, a fundamental-cut crystal is used, and for frequencies above 20 MHz, an overtone crystal is used. When using the overtone crystal, C5 is replaced by a short and C6 is reduced to 10 pF. The load capacitance of the oscillator circuit is 32 pF and must be specified when ordering crystals.*

Ordinary inductors can be used for L1, L2 and L3, but I recommend toroids. Silver mica capacitors are used for C1-C6, as well as the 20 pF, 100 pF, and the 10 pF capacitors in the circuit. It's not advisable to replace these capacitors with ceramic types. The i-f transformers come as a set from Radio Shack (part number 273-1383). T1 has the gray core; T2 the white core. The other transformer and the oscillator coil are not used. If you substitute transformers, make certain they have the same pin connections if you use the printed circuit. T3, the output transformer, is a Radio Shack item, part number 273-1381. Almost any germanium diode can be used for D1 in place of the 1N277, and many silicon npn transistors can be used for Q3.

Construction practices are not critical. Keep lead lengths short and use sufficient bypassing on the power buses. Oscillator tank coil L3 should be isolated from the rf tuned circuits to prevent desensitizing the rf amplifier. There are no special handling precautions for Q1, since it is a protected gate type.

A printed circuit layout is shown in figs. 4 and 5. This is a double-sided board, which uses one side as the

**table 1. Receiver front-end component values for receiving WWV, WWVH, and CHU.**

| station | freq (MHz) | crystal freq (MHz) | C1 value (pF) | C2 value (pF) | C3,C4 values (pF) | C5 value (pF) | C6 value (pF) | L1&L2 turns | L1&L2 AWG | L1&L2 (mm) | L3 AWG | L3 (mm) | coil cores† |
|---|---|---|---|---|---|---|---|---|---|---|---|---|---|
| WWV WWVH | 2.50 | 2.955 | 300 | 820 | 220 | 30 | 150 | 66 | 32 | (0.2) | 32 | (0.2) | T37-2 |
| WWV WWVH | 5.00 | 5.455 | 120 | 680 | 100 | 30 | 150 | 49 | 32 | (0.2) | 32 | (0.2) | T37-2 |
| WWV WWVH | 10.00 | 10.455 | 56 | 330 | 47 | 30 | 150 | 40 | 32 | (0.2) | 32 | (0.2) | T25-2 |
| WWV WWVH | 15.00 | 15.455 | 33 | 330 | 30 | 30 | 150 | 37 | 30 | (0.25) | 30 | (0.25) | T25-6 |
| WWV WWVH | 20.00 | 20.455* | 30 | 330 | 27 | short | 10 | 29 | 32 | (0.2) | 32 | (0.2) | T25-6 |
| WWV WWVH | 25.00 | 25.455* | 24 | 300 | 22 | short | 10 | 26 | 32 | (0.2) | 32 | (0.2) | T25-6 |
| CHU | 3.33 | 3.785 | 300 | 820 | 220 | 30 | 150 | 50 | 30 | (0.25) | 30 | (0.25) | T37-2 |
| CHU | 7.34 | 7.795 | 68 | 350 | 56 | 30 | 150 | 44 | 32 | (0.2) | 32 | (0.2) | T37-2 |
| CHU | 14.67 | 15.125 | 33 | 330 | 30 | 30 | 150 | 36 | 32 | (0.2) | 32 | (0.2) | T25-6 |

*Overtone crystal

†Available from Amidon Associates, 12033 Otsego St., North Hollywood, California 91607

*Available from JAN Crystals, 2400 Crystal Drive, Fort Myers, Florida 33901.

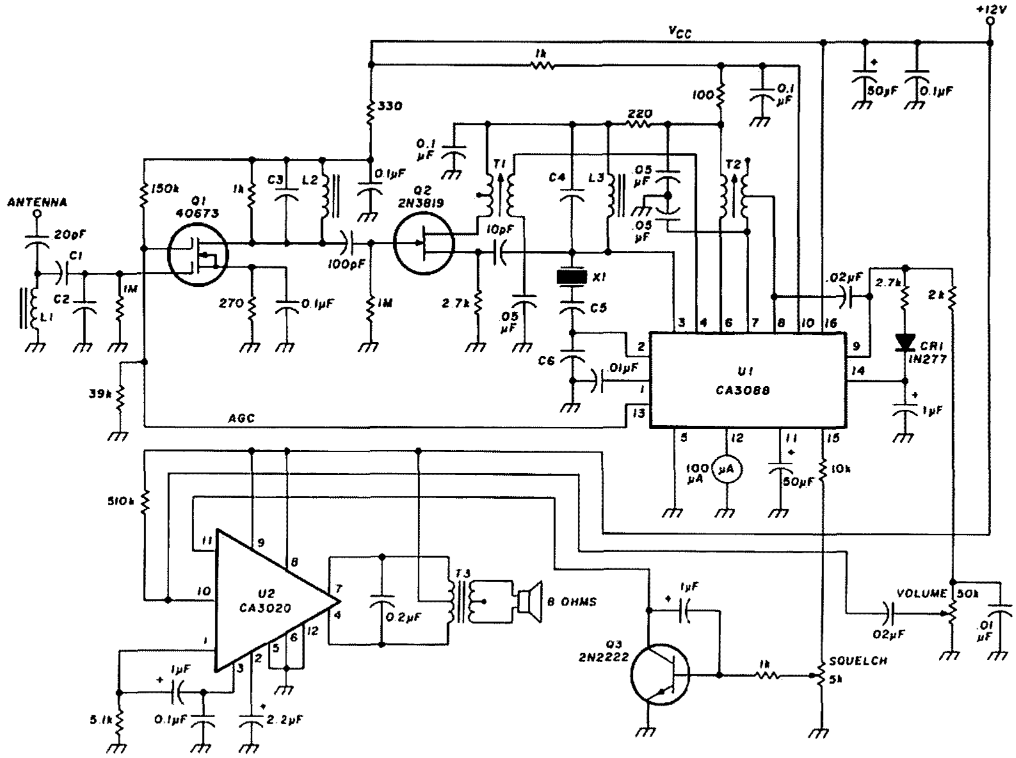
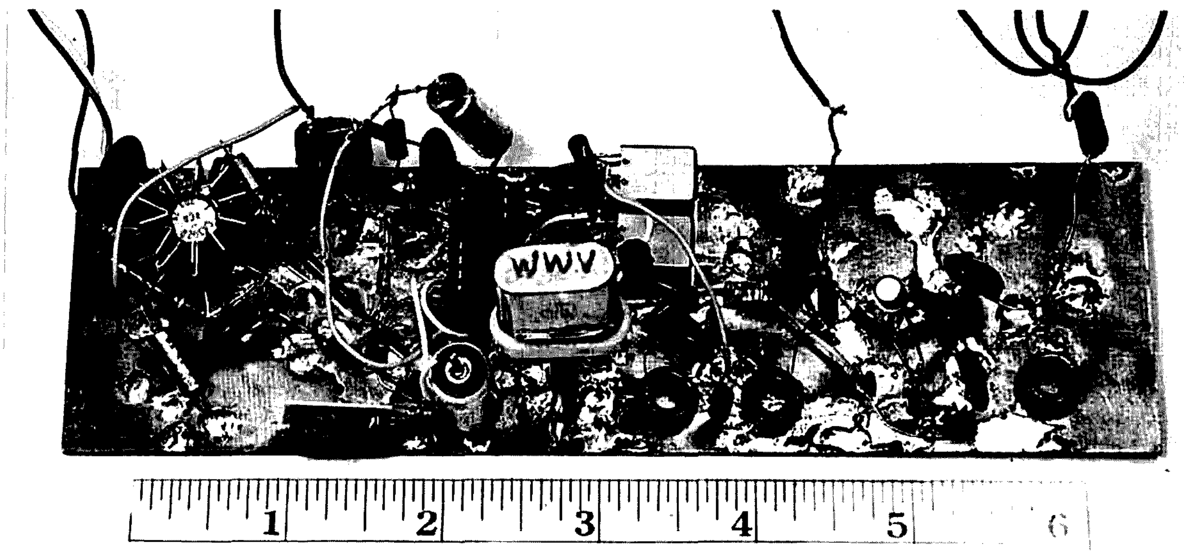



fig. 3. Receiver schematic. Broadband toroids are used instead of slug-tuned coils in the rf, mixer, and local-oscillator to reduce circuit complexity and increase stability.

ground plane. Some of the components are soldered to both sides of the board, so to avoid burned fingers and incinerated components, some planning is necessary before assembling the PC board. Solder the components that are connected to the ground plane first. Solder the shortest components first, then solder components in order of increasing height. This will prevent the difficulty of soldering a component to the ground plane that is surrounded by taller components already inserted.




Parts placement in the prototype version of the 10-MHz WWV receiver. Power requirements are 12 volts at 100 mA. For portable work, alkaline penlight cells will provide hours of service.

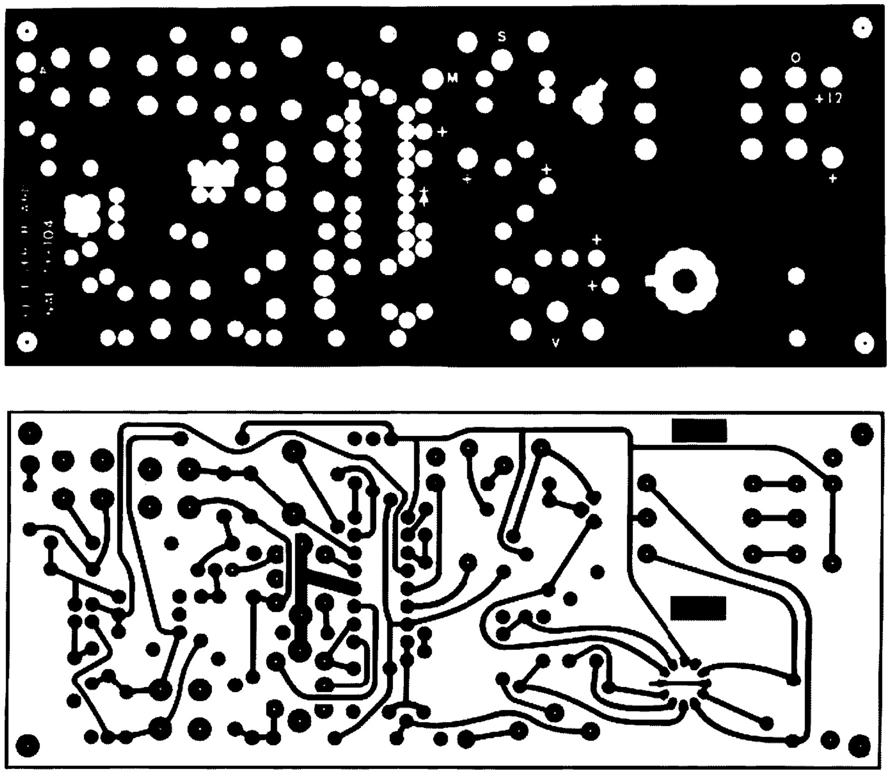


fig. 4 Top and bottom of the receiver PC board. A double-sided layout is used, with some of the components soldered to both sides of the board.

There is one jumper wire on the circuit board, which connects pins 5, 6, and 12 of the CA3020 to the ground plane. Solder a piece of bus wire through the hole from the ground plane to the pad underneath. Soldering the CA3020 into the board can be tricky. The pads are very small and overheating may cause them to lift. One method that works is to wrap a short piece of no. 18 (1mm) bus wire around the tip of the soldering iron and use this as a miniature tip to solder the IC. Plastic spacer pads should be used under Q1 and Q3 to prevent their cases from shorting to the ground plane. Finally, use a heat sink on the CA3020, since it dissipates enough power to get warm.

After assembly, a dc check of the circuit should be made. Apply 12 volts to the circuit and measure the current drawn. The current should be around 65 milliamperes. If it's much higher than this, remove power immediately and check the wiring.

## alignment and operation

If the toroidal inductors are used, the receiver front end is adjustment-free and alignment is easy. Loosely

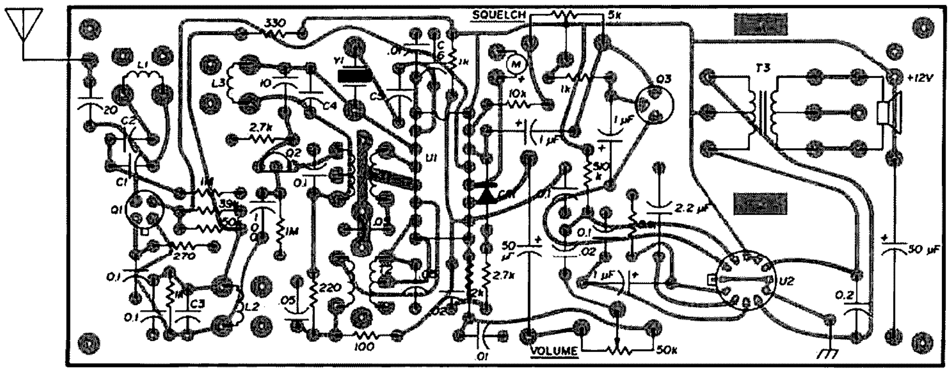



fig. 5. The receiver PC board with loading diagram for parts placement.

couple a generator to the antenna input and set it for the operating frequency. Adjust T2 for maximum output from the speaker or the tuning meter. Repeat for T1, and alignment is complete. In the case of the 10-MHz prototype, an antenna was connected to the reciever and the signal from WWV itself was used instead of a generator.

To align a receiver with slug-tuned coils, pull out the crystal and couple the generator into Q2 drain. Set the generator for 455 kHz, tune T2 for maximum, then tune T1 for maximum. Disconnect the generator and insert the crystal. Adjust L3 until an increase in background noise is heard in the speaker. Couple the generator into the antenna and set it for the operating frequency. Adjust L2 for maximum signal, then adjust L1 for maximum. Disconnect the generator and you are ready to go.

If stability problems develop with slug-tuned coils, it may be necessary to decrease the value of the 1k resistor across the rf stage output tank circuit. This lowers the tuned-circuit Q, so if instability remains, the only alternative may be to use the toroids.

A few feet of hookup wire is all that was needed for an antenna on the prototype. Receiver location, propagation conditions, and operating frequency will affect the amount of antenna required to get a good signal. Grounding also improves signal strength when the receiver is not used for portable work.

Any small, reasonably regulated power supply that can deliver 12 volts at 100 mA is suitable for powering the receiver. It's important that the voltage not exceed 12 volts, since the CA3020 rating will be exceeded. Batteries also work well, and some alkaline penlight cells will give hours of service.

## performance

The prototype receiver has been very satisfactory in meeting all of the initial requirements. A receiver was also constructed for the 5-MHz WWV carrier and worked very well during late evening hours when the 10-MHz signal was weak. A bandswitching arrangement would be an excellent idea. The other frequencies have not been used on a prototype receiver, but performance should be equally good if the component values given in table 1 are used.

Use of the ICs in the design reduced size and cost of the project tremendously. The CA3088 is very impressive in its performance, flexibility, and cost. It won't be surprising if other excellent designs result from its use.

### references


1. *RCA Linear Integrated Circuits and MOS Devices Data Book,* RCA publication SSD-201B, 1974, page 446.
2. *Radio Amateur's Handbook,* ARRL, Newington, Connecticut, 51st Edition, 1974, page 246.
3. *RCA Linear Integrated Circuits and MOS Devices Data Book,* "Application Notes," RCA publication SSD-202B, 1974, page 318.
4. *RCA Linear Integrated Circuits Manual,* RCA publication IC-42, 1970, page 226.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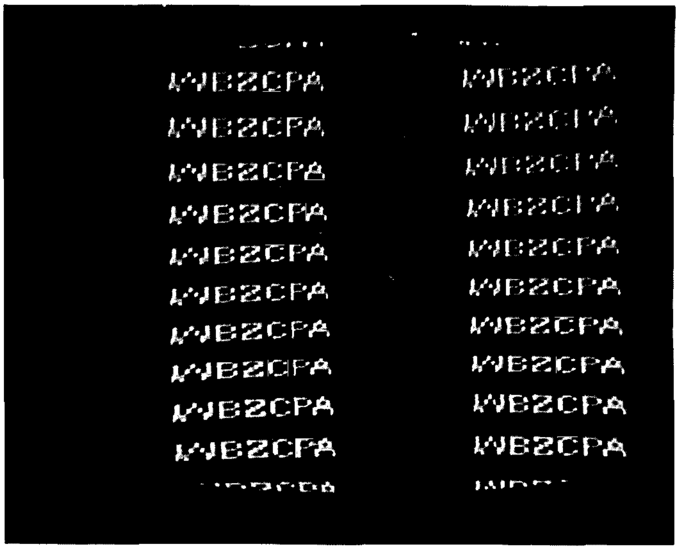


# amateur television callsign generator

## Callsign generator for amateur TV uses programmable memory to produce a professional-looking identification display without a camera

With the widespread availability of programmable ROMs (PROMs) it has become possible to build an ATV callsign generator with less effort than ever before. Such a generator allows you to identify your ATV station without the use of a camera. Carefully lettered signs that somehow don't look professional are a thing of the past at WB2CPA. My homemade RS-170 sync generator and vertical interval switcher has an *identify* switch that superimposes my call on the outgoing video.

The heart of my character generator is an ROM (Read Only Memory), and I used the readily available 32 x 8 PROM. The designation 32 x 8 means that the PROM has 8 outputs and 5 inputs ($32 = 2^5$). For each of the 32 possible states of the 5 inputs an 8-bit word will appear at the output. I used that fact to produce the TV image.

The first step in preparing your custom ROM is to make a *truth table* (**fig. 1**) representing your call. On a piece of graph paper mark 32 rows of 8 lines each. Now draw your call letters into these 256 boxes, striving for best appearance and fit.

Amateurs with two-by-three calls such as mine will find that only 7 of 8 lines can be filled to obtain well-proportioned letters. The graphic information (white spaces are logic 1 and dark spaces are logic 0) must now be placed into a ROM. There are several ways to do this. *Popular Electronics*[1] had plans for a programmer, and lists a vendor for the blank PROM.* Or you can have it done by any of several distributors who will program a memory to your truth table at reasonable cost.†

Five other ICs are needed to complete your callsign generator. The complete ROM must have all 32 locations addressed at least once per TV line and its 8 outputs sampled sequentially, advancing to the next after the end of each line on the screen (**fig. 2**).

*Signetics 8223, James Electronics, P.O. Box 822, Belmont, California 94002

†Solid State Systems, Box 617, Columbia, Missouri 65201.

By **Jerry Pulice, WB2CPA,** 143 Gibson Avenue, Staten Island, New York 10308

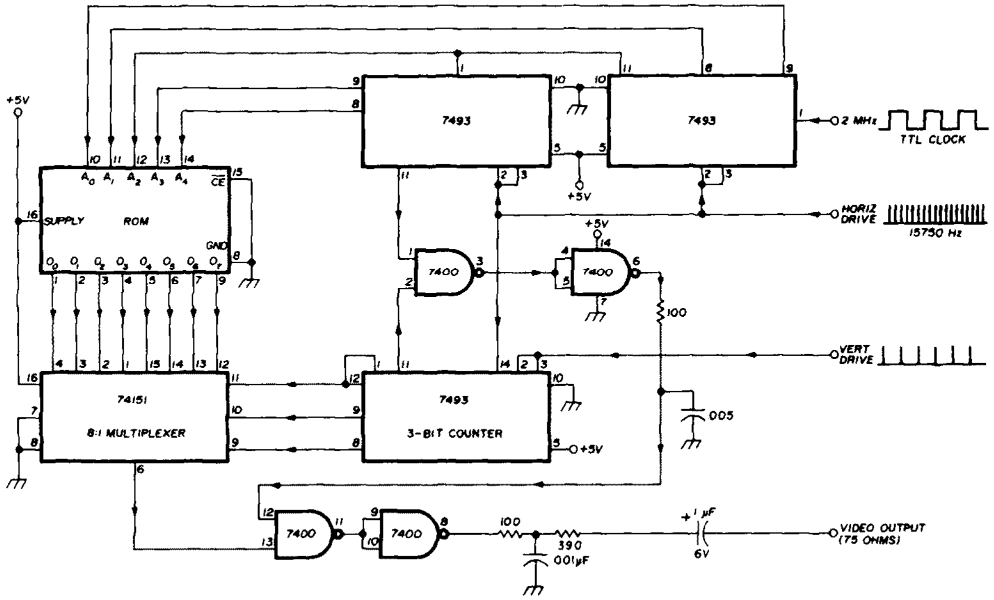


fig. 2. Logic diagram for the ATV callsign generator. Complete circuit operation is described in the text. The ROM pin connections shown here are valid for the AMI 27508/27509, 82523/825123, MM5330/MM5331, HPROM 8256, or IM5600/5610.

I used a pair of 7493 binary counters to address the 32 words in the ROM. The rate that the 7493s are clocked affects the length of the characters on the screen. If you have 2- to 3-MHz TTL pulses somewhere in your sync generator (as I have) they can be used to clock the counter. If not, use the TTL oscillator shown in fig. 3.

The 74151 multiplexer must advance to the next ROM output once per scan line. Its 3-bit counter is clocked by horizontal drive pulses from the sync generator. The characters on the screen should appear in the same position on each field in order to remain stationary. Therefore, reset pulses are applied to the counters. Positive-going horizontal drive pulses reset the 5-bit word counters and positive-going vertical drive pulses reset the 3-bit line counters. For the display to be stable, these pulses must come from the same sync generator.

Video output of either polarity can then be taken

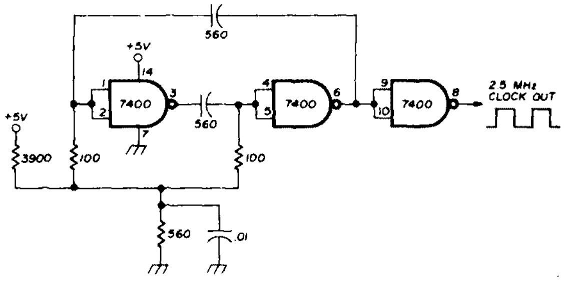


fig. 3. Suggested TTL oscillator circuit for use with the ATV callsign generator.

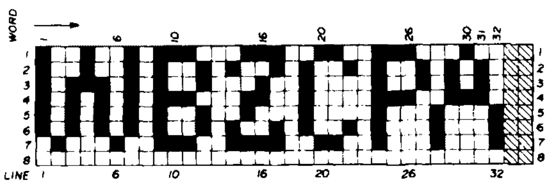


fig. 1. Truth table for the ATV callsign generator showing graphic layout of WB2CPA's callsign. Available space can be layed out for any two-by-three amateur callsign, or for shorter calls.

from the 74151 multiplexer. Extra bits from the word and line counters are used to gate the video *on* only 50 percent of the time. This puts a border around the letters to increase clarity.

To add the video to an existing 75-ohm source, merely connect the vertical and horizontal drive *from* the source and couple it in parallel with the TTL output through the RC network shown in fig. 2.

It is illegal to add the suffix "TV" to an amateur station callsign, so I added a touch to my generator that uses the fact that my PROMs have tri-state outputs. I have two ROMs in parallel and alternate between them by using their chip-enable inputs. In this way the image *alternates* between WB2CPA and TV. This impresses the heck out of visitors to the shack!

## reference

1. Robert D. Pascoe, "How to Program Read-Only Memories," *Popular Electronics*, July, 1975, page 27.

**ham radio**

# how to make custom capacitors for your homebrew projects

## Construction details for making your own low-inductance capacitors from PC-board stock

Recently the need for some low-value, low-inductance capacitors led me on a merry chase through local radio parts outlets. The capacitors I needed were either not in stock or too expensive. I'm used to building almost everything I need, as are most vhf men, but in this case it took a bit more thought.

After several dismal attempts at making capacitors with casting resin, thin brass sheet, paper, and glue, the light suddenly snapped on – printed-circuit board! PC boards are available in a variety of thicknesses and dielectric material at reasonable prices. I chose the epoxy paper type (double-side board) and estimated the dielectric constant at 3.3. Material thickness was 0.0625 inch (1.59mm).

### computations

After cutting several pieces of PC-board material with a nibbling tool, I computed the capacitance values (approximately, but I was surprisingly close!) and verified them on a "Mickey Miker." I attribute any small errors to lack of knowledge of the exact dielectric constant of the PC-board material. The dimensions and

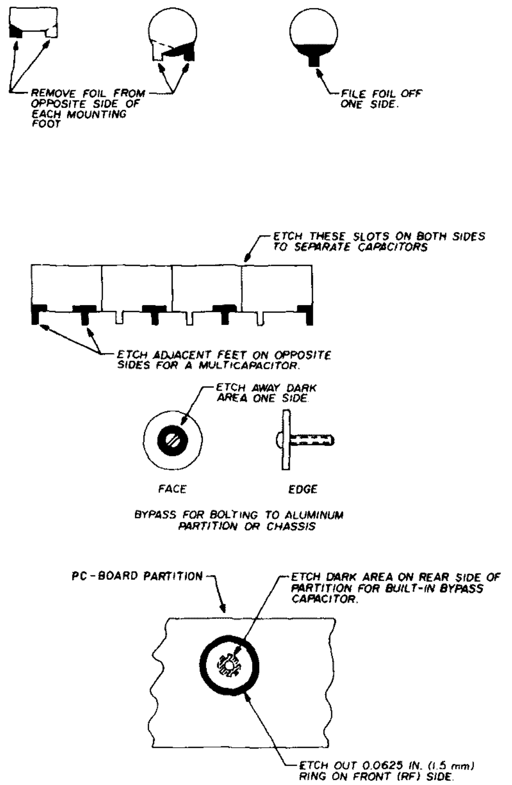


**fig. 1. Construction details for making your custom capacitors from PC-board material. All have low inductance and are suitable for use with printed circuits. Table 2 gives capacitance values for various diameters, using double-sided epoxy paper type PC-board material 0.0625 in. (1.59mm) thick.**

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table 1. Matrix showing capacitance values for square or rectangular shapes. Accuracies are sufficient for experimental amateur work.

| dimension A, in. (mm) | dimension B, in. (mm) 0.25(6.5) | 0.5(12.5) | 0.75(19) | 1.0(25.5) | 1.25(31.8) | 1.5(38) | 1.75(44.5) | 2.0(51) |
|---|---|---|---|---|---|---|---|---|
| | capacitance, pF | | | | | | | |
| 0.25(6.5) | 0.7 | 1.5 | 2 | 3 | 4 | 4.5 | 5 | 6 |
| 0.5(12.5) | 1.5 | 3 | 4.5 | 6 | 7 | 9 | 10 | 12 |
| 0.75(19) | 2 | 4.5 | 7 | 9 | 11 | 13 | 15 | 18 |
| 1.0(25.5) | 3 | 6 | 9 | 12 | 15 | 18 | 21 | 24 |
| 1.25(31.8) | 4 | 7 | 11 | 15 | 18 | 22 | 26 | 30 |
| 1.5(38) | 4.5 | 9 | 13 | 18 | 22 | 27 | 31 | 35 |
| 1.75(44.5) | 5 | 10 | 15 | 21 | 26 | 31 | 36 | 41 |
| 2.0(51) | 6 | 12 | 18 | 24 | 30 | 35 | 41 | 47 |

values in the tables are sufficient for amateur purposes anyway, since our work is mostly cut and try. The equation used for computing capacitance is:

$$C = 0.224 \frac{KA}{d(n-1)} \quad (1)$$

where

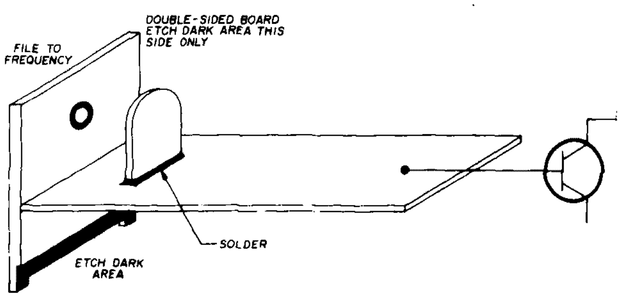
$C$ = capacitance (pF)
$K$ = dielectric constant
$A$ = area of one plate (square inches)
$d$ = spacing of plates (PC-board thickness, inches)
$n$ = number of plates

For using metric equivalents **eq. 1** becomes:

$$C = 0.00882 \frac{KA}{d(n-1)} \quad (2)$$

where

$A$ = area of one plate ($mm^2$)
$d$ = spacing of plates (PC-board thickness, mm)



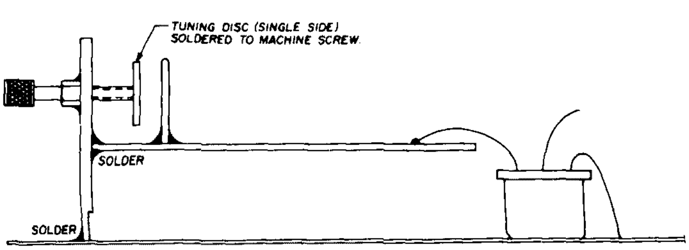




fig. 2. A capacitor and stripline inductor combination for microwave applications.

## Examples

Using $d = 0.0625$ inch, $A = 2$ $inch^2$ and $K = 3.3$:

$$C = 0.224 \frac{3.3 \times 2.0}{0.0625 \times 1} = 23.65 \; pF$$

Using $d = 1.59$ mm, $A = 1290$ $mm^2$ and $K = 3.3$:

$$C = 0.00882 \frac{3.3 \times 1290}{1.59 \times 1} = 23.61 \; pF$$

The small difference in the two examples is due to roundoff errors in making the metric conversion. **Table 1** is a matrix allowing you to choose capacitance values for *various plate dimensions, using square or rectangular* shapes, and **table 2** is a handy reference for round capacitors.

table 2. Capacitance values for round capacitors made as shown in fig. 1.

| dimension D, in. | (mm) | capacitance, pF |
|---|---|---|
| 0.25 | (6.5) | 0.58 |
| 0.3125 | (8.0) | 0.98 |
| 0.375 | (9.5) | 1.30 |
| 0.4375 | (11.0) | 1.77 |
| 0.5 | (12.5) | 2.32 |
| 0.5625 | (14.5) | 2.93 |
| 0.625 | (16.0) | 3.62 |
| 0.6875 | (17.5) | 4.38 |
| 0.75 | (19.0) | 5.22 |
| 0.8125 | (20.5) | 6.09 |
| 0.875 | (22.0) | 7.11 |
| 0.9375 | (24.0) | 8.15 |
| 1.0 | (25.5) | 9.28 |

## construction

The capacitors in **fig. 1** have very low inductance and are suitable for mounting on printed circuits. An interesting point is that to achieve the perfect value for your needs you can sand or file an edge a bit at a time, slowly reducing the capacitance. **Fig. 2** shows how to make a PC-board capacitor and stripline inductor combination for microwave circuits. No doubt with a little thought many other capacitor applications will come to mind.

**ham radio**

# isolating parallel currents in rf amplifiers

## Construction techniques improving amplifiers stability using toroid cores and ferrite beads

**Problems with radio-frequency amplifiers** first occurred when I decided to build them on a metal chassis, which was then put inside a metal box. I also used large areas of the chassis as part of the conductors for the input and output tuned circuits. This construction technique resulted in a very low impedance path for the conduction of parallel currents that find their way into the amplifier from the antenna. How I solved instability problems in an rf amplifier resulting from these parallel currents is the subject of this article.

### parallel currents

Out-of-band signals propagating down the antenna transmission line as parallel currents are not rejected by the amplifier tuned circuits. These currents pass straight on to the mixer either directly or by being induced into amplifier tuned circuits by way of the chassis. A typical path taken by the parallel currents in the input section of an rf amplifier is shown in **fig. 1**. These currents mix with strong in-band signals causing intermodulation products that produce distortion.[1]

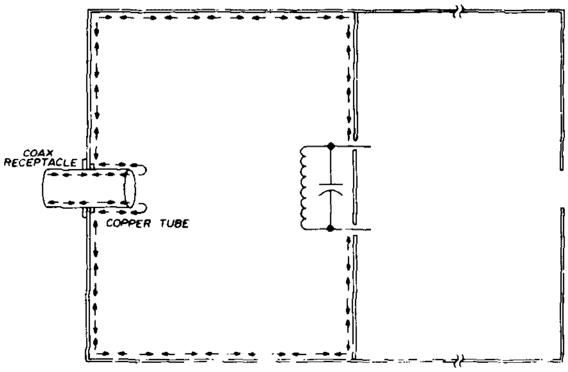



fig. 1. Path of parallel currents in the input section of an rf amplifier. The currents are induced in the antenna, travel down the braid of the transmission-line coax cable, and find their way into the amplifier to cause IMD and feedback problems.

Fig. 2 shows how parallel currents are set up on the *antenna and coax feed system*. The currents are apparently at a much higher level than is generally realized,

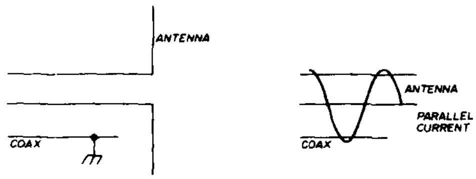



fig. 2. How parallel currents are set up on the antenna and feed line. The antenna does not balance out these currents. They are in phase on the antenna element.

especially when the antenna is mounted on a large metal object such as a ship, or when mounted on a building containing extensive metal conduits and electrical wiring. The parallel currents travel along these conducting devices then are induced into the outer side of the coax cable, travel up it to the antenna and then down both inner conductor and inside of the coax braid. The antenna does not balance out these currents; they *are in phase* on both sides of the antenna element. Such currents also cause increased levels of internal feedback in an rf amplifier, which make it difficult to obtain good neutralizing adjustments.

### isolation method

The schematic of fig. 3 shows methods I used to solve the problems caused by parallel currents. Input and output link coils and lengths of copper tubing on the coax connectors separate true antenna currents from parallel currents. A high impedance to the parallel currents is provided by the toroid cores placed over the copper tubing on input and output connectors, so the

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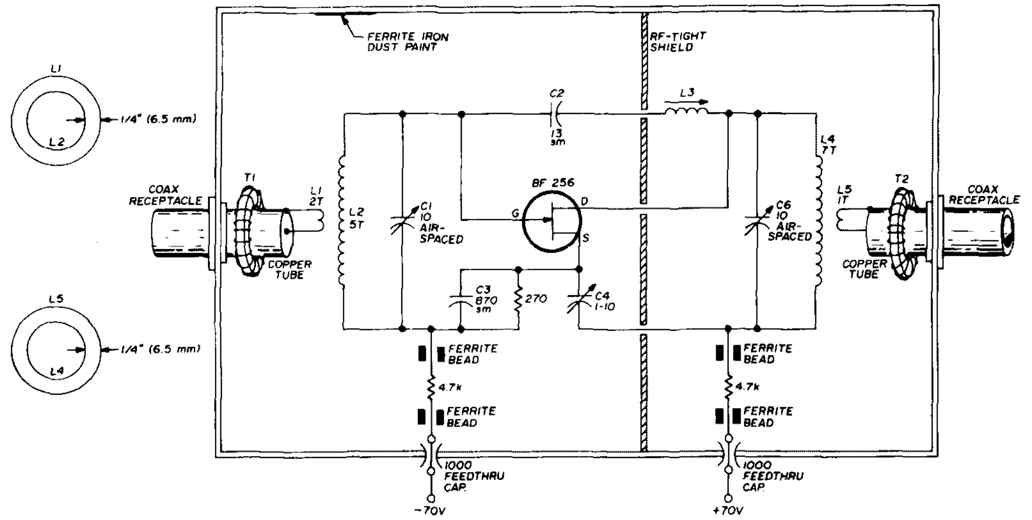


**fig. 3. Rf amplifier with improved isolation of parallel currents between input and output. Sketches at top are end views showing how input and output links are coupled to the amplifier-tuned circuits. The T-68-10 toroid holes were enlarged slightly to slip over the copper tubing soldered to the coax receptacles.**

currents can't reach the following stage. Ferrite beads on the wiring carrying input voltage to the amplifier provide further isolation. An rf-tight shield placed across the center of the box isolates input and output circuits. Finally, a coat of ferrite-dust paint was applied to the inside of the box.

The amplifier shown in **fig. 3** has been fully tested on the air. The only thing that doesn't appear to improve matters is the ferrite dust painted on the inside of the box. The toroids improved amplifier stability and neutralizing adjustments.

### reference


1. James R. Fisk, W1DTY, "Receiver Noise Figure, Sensitivity and Dynamic Range — What the Numbers Mean," *ham radio*, October, 1975, page 18.


**ham radio**

# loops and reflectors

ZL1BN's article on corner-fed loop antennas for low and medium frequency use inspired some thoughts of improving my own forty-meter performance, especially for long-haul and DX work.* I immediately built a loop for forty meters and suspended it from a thirty-foot mast attached to the center support of my neighbor's back yard chain link fence. This happened to be a fortuitous choice for a couple of reasons, since my neighbor — W2OSY — and I have been hamming together for a number of years, and his fence has been used on a shared basis for many antenna projects. The second reason is a bit more technical, because the fence turned out to be an excellent linear reflector element for the loop.

In order that the horizontal element of the triangle be raised sufficiently above ground to prevent decapitating my son when he mowed the lawn, I had to slant the antenna by pulling the bottom portion horizontally away from the fence. This was done with light nylon line attached to each lower corner, and resulted in the bottom element of the loop being about ten feet off the ground. It turned out that the median distance from the loop center to the top of the fence was about 0.15 wavelength on forty; not bad for a reflector. I fed the antenna with RG58 coaxial cable at the corner closest to the shack, and tuned my transmitter to check on the swr. Surprise! The swr was below 2:1 over the entire CW portion of the band. On-the-air checks showed that my signals were excellent into eastern Europe and the Mediterranean, as well as into western Africa. Results in the other direction (to the west) showed a very pronounced front-to-back ratio, with only very high-angle, close-in signals being received. No West Coast U.S. signals were heard during many hours of late evening operation, contrary to my usual experience.

I'm sure that a triangular tuned reflector added to the loop would show even better performance; but as it is, this antenna into Europe out-performs any forty meter antenna I've ever used, so I want to say thanks to Barry Kirkwood, ZL1BN, for a most timely and satisfactory aid to my DX efforts. The corner-fed system in particular makes me and my transmitter very happy, because I don't need a tuner anymore. Now, before snow flies, I'm planning a delta loop for 75 and 80 meters at my new QTH. Sure wish I had that fence in my back yard here, too!

Jim Gray, W2EUQ

*April, 1976, *ham radio*

# silver plating

## made easy

**Silver plating is neither difficult nor expensive — this article shows you how**

The controversy as to the merit of silver plating has existed for many years and from time to time is still debated on the ham bands. The experts are nearly unanimous in recommending silver plating, particularly on vhf and microwave conducting surfaces.

Aside from the improvement in electrical performance that is generally claimed, there is, in my opinion, a more important reason for silver plating – preservation. It would be interesting to wind a pair of two-meter coils from bare copper wire, silver plate one, and leave the other unplated. A comparison of their performance should show little or no noticeable difference. However, repeat the experiment two years, or even six months, later and the results may be entirely different. The *silver-plated coil would* very likely show superior performance, and it would look better.

### materials

Brass and copper are ideal raw materials frequently used for homebrew projects. Both are easy to work and they lend themselves to silver plating. Thin sheet stock and various sizes of tubing can be purchased in many hobby shops. Scrap brass can often be purchased in a wide variety of sizes from junk shops, and at reasonable prices. Flashing copper may be available from a local roofing contractor or hardware store. It is also possible to purchase brass or copper from steel supply houses, but the cost is higher and there are other constraints, such as *minimum order* quantities which are in excess of the average amateur's use.

**Table 1. Relative resistivities of various metals used in amateur projects.**

| | |
|---|---|
| Silver | 0.94 |
| Copper | 1.00 |
| Aluminum | 1.60 |
| Zinc | 3.40 |
| Brass | 3.70-4.90 |
| Nickel | 5.10 |
| Tin | 6.70 |
| Lead | 12.80 |

### silver plating

Aside from the silver plating controversy, there are theoretical considerations which show that circuit losses increase with frequency. One way to reduce losses is to use silver plating. Because of a phenomenon called skin-effect,[1] rf current tends to concentrate near the surface of a conductor. This is caused by the way the magnetic field, produced by the current flowing in the wire, is distributed in and about the conductor.[2] The inductance of the wire is greater at its center, thus providing an easier path for the current near the outside surface.

Skin depth is defined as that distance below the surface of the conductor where the current density has dropped to about 37 per cent of the density at the surface.[3] For example, a straight round wire at 144 MHz has a skin depth of approximately 0.216 mils (0.0055mm). As the frequency increases, the skin depth decreases according to the following relationship:

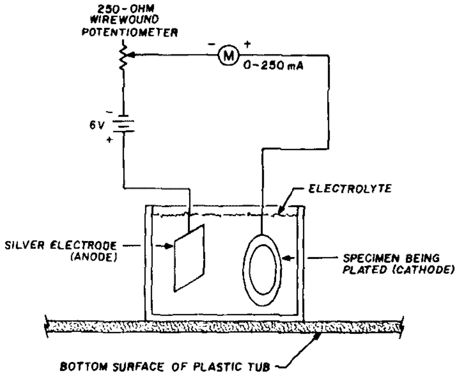


**fig. 1. Hookup for silver plating. Recipe for the silver-plating solution (electrolyte) is given in table 2.**

$$d = 2.59 \sqrt{\frac{1}{f}} \text{ mils}$$

where $f$ is in MHz. At 220 MHz, skin depth is about 0.175 mils (0.0044mm). At 432 MHz, it is only about 0.125 mils (0.0032mm).

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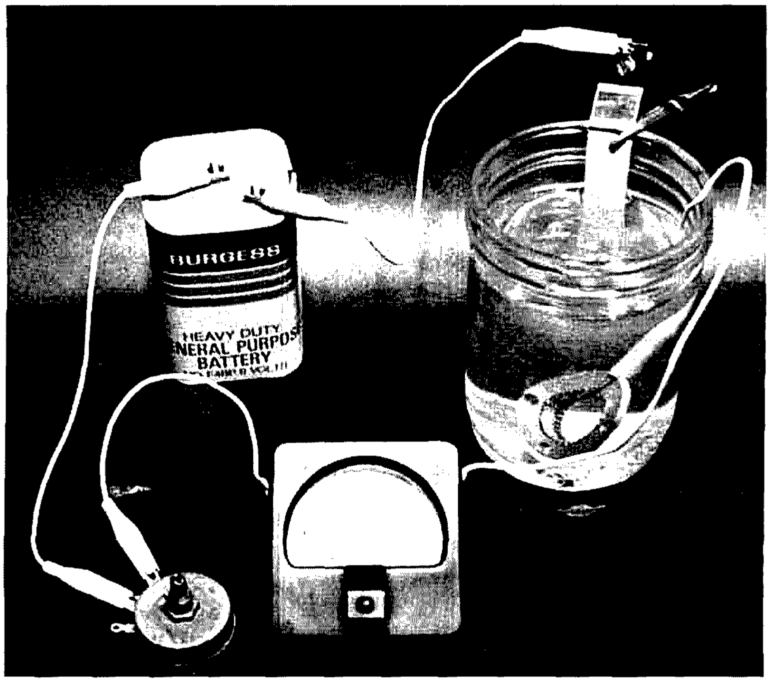


Table 1 lists the relative resistivities of various metals likely to be used for amateur construction projects.[4] While clean copper is only slightly more resistive than silver, the difference becomes greater as the copper oxidizes. Note that there is an obvious need to silver plate brass.

The types and quantities of chemicals needed to mix a gallon of silver-plating solution are given in table 2. The chemicals should be poured slowly while stirring into a wide-mouth jar containing about two-thirds of a gallon (2.5 liters) of distilled water. Add the chemicals in the order shown in the table. Mixing should be done out-of-doors where there is adequate ventilation since a small amount of cyanide gas will be given off in the mixing process. *Caution: Cyanide gas is poisonous — avoid contact of the solution with open sores and avoid prolonged breathing of the vapors.*

After the plating solution has been prepared, it should be stored in a one-gallon (4 liter) jug with a narrow neck and a screw cap. Pour the plating solution into the gallon jug with the aid of a small aluminum funnel. Add distilled water to bring the volume up to a full gallon (3.8 liters). Be sure to store the solution out of reach of children.

## current density

The current density recommended by experienced platers is between 5 and 15 amperes per square foot (5 to 16 mA per square cm) of plating surface. On this basis, a piece of 0.032-inch (0.8mm) thick brass plate, one inch (25mm) square requires a plating current of between 35 and 105 milliamperes, if the edges are neglected. The thickness of the deposited silver depends on the amount of current and the plating time. Experience has shown that for hobby uses, as a general rule, it is seldom necessary to exceed a few hundred milliamperes of current for more than a half hour.

## plating thickness

The electrochemical equivalent for silver is 4.0255 grams per ampere-hour.[5] This means one gram will be plated in one hour at a current of 248 milliamperes. A

one-ounce spoon of electrolytic silver weighs approximately 28 grams. Since its specific gravity is 10.5, its volume is 28/10.5 or 2.7 cc. When plating for a half hour at 100 milliamperes, 0.201 grams is deposited, which is equivalent to 0.019 cc (0.201/10.5 = 0.019). The silver spoon will be completely used up after 140 such plating sessions.

The thickness of silver deposited on a thin, one-inch-square (6.4 square cm) plate in a half hour at 100 milliamperes is calculated as follows:

$$t = \frac{0.061V}{A}$$

where: $t$ = thickness (mils)
$V$ = volume deposited (cc)
$A$ = surface area (square inches)

$$t = \frac{0.061 \cdot 0.019}{2 \cdot 1^2} = 0.580 \; mil$$

The figure 2 appears in the denominator because the 1-inch square plate has two sides. This is a reasonable thickness for hobby work, and it exceeds the skin depth at frequencies above 100 MHz. At lower frequencies, the silver serves more as a protective layer than a low-loss plating.

The chemical plating process[6] involves an electrolyte ionized by the applied voltage. Negative ions drift from the cathode to the silver anode. The silver breaks up into positive ions which go into solution. The free electrons travel over the external circuit to the cathode to combine with silver ions coming out of solution, thus forming metallic silver. The silver anode is gradually dissolved and conveyed to the cathode. The amount and composition of the electrolytic does not change in this process. Except for a very gradual contamination from other causes, the electrolyte has indefinite life.

**Table 2. Silver-plating solution recipe.**

| | |
|---|---|
| Silver cyanide | 4.8 oz (136 grams) |
| Potassium cyanide | 8.0 oz (226 grams) |
| Potassium carbonate | 2.0 oz (57 grams) |
| Distilled water | 1.0 gal (3.8 liters) |

## preparation for plating

The most important word of advice for the would-be plater is cleanliness. Not only must the specimen to be plated be spotlessly clean, care should be taken to avoid contaminating the electrolyte with cleaning agents. Two plastic pails should be used for cleaning purposes. Both should be located in one compartment of a double sanitary tub. The plating solution should be located in the other compartment. One pail should be filled with cold tap water to which is added a quarter-cup (2 ounces or 59cc) of liquid *All* detergent, or equivalent. This pail will serve for cleaning the specimen to be plated. The other pail should be located under the cold-water tap, with the water running, and used for rinsing purposes. This will help to avoid contaminating the electrolyte.

Very fine steel wool such as 000 grade, saturated in detergent, makes an excellent scrubber, especially for brass which is more difficult to clean than copper. After scrubbing, rinse the specimen and submerge it in the electrolyte, but out of contact with the silver anode. Adjust the current to 30 milliamperes and remove the specimen after one or two minutes of plating.

Rinse and examine the specimen for chalky regions which identify areas not initially clean. Repeat the scrubbing process, wash, and return the specimen to the electrolyte. After another few minutes, repeat the inspection process. Seldom will the cleaning process have to be repeated more than twice. A final wash and continued plating for 15 to 30 minutes at a current of 100 milliamperes will complete the job. While larger specimens require longer plating times, the process can be speeded up if the current is increased. However, if the plating is done too rapidly, a chalky finish may appear which rubs off when scrubbed with steel wool and detergent. It is prefereable to plate at a slow rate. It is a good idea to remove the specimen after ten or fifteen minutes to assure that the plating speed is not too fast. Also, while the specimen is in the final plating process, change its position from time to time relative to the anode to assure a more uniform plating.

The last step is to wash and soak the piece in the detergent pail for ten minutes or more. Run tap water into the pail until the water is clear. This washes away traces of electrolyte which might otherwise leave spots. A final word of advice: Be sure your hands are clean when handling the freshly plated specimen; skin oils leave finger prints. Dry the plated specimen as completely as possible with a soft cloth to avoid water marks.

## preserving the shine

While your hands are still clean, use masking tape to cover those areas of the specimen which will be used for electrical contact with other parts. Then spray the specimen on both sides with a thin coating of clear enamel. An inexpensive, non-toxic clear enamel called *One Coat*, is distributed by Korvettes, and does a fine job of protecting the silver surface from discoloration due to oxidation. If the instructions given here are carefully followed, the silver-plated specimen will be a thing of beauty, with a bright silver coat.

## references

1. K.R. Sturley, *Radio Receiver Design*, Part I, John Wiley and Sons, Inc., London, 1942, page 131.
2. F.E. Terman, *Radio Engineering*, McGraw-Hill, New York, 1937, page 35.
3. *Reference Data for Radio Engineers*, 4th edition, ITT, Strafford Press, New York, page 128.
4. *The Radio Amateur's Handbook*, ARRL, Newington, Connecticut, 1977, page 18.
5. Charles Hodgman, editor-in-chief, *Handbook of Chemistry and Physics*, 23rd edition, Chemical Rubber Printing Co., Cleveland, Ohio, page 1525.
6. W. H. Timbie, *Elements of Electricity*, 4th edition, John Wiley and Sons, New York, 1959, page 295.

**ham radio**

# bandspreading techniques
## for resonant circuits

How to calculate capacitance values for tuning the desired bandwidth in variable-frequency oscillators, bandpass filters, and other tuned circuits

Adding series and parallel capacitors to a variable tuning capacitor are common techniques in bandspreading a vfo or other tuned circuit. However, very little information has been published on finding the correct values of the added capacitance. The calculations given here are relatively simple and assure the designer of quickly finding the correct values.

An inexpensive pocket calculator with a square-root function will improve accuracy and help keep track of the decimal point. Test equipment is needed only if capacitance values are unknown, and a grid dipper with a few standard, known-value capacitors will do the job nicely.

All calculations are simple algebra, so don't be worried about the technique. All you have to do is keep the decimal points in their proper places and plug in the correct values where they're called for.

### basis for the technique

The key to this technique is based on the ratio of maximum to minimum capacitance and frequency. The basic resonance formula,

$$f^2 = \frac{25330.3}{LC} \qquad (1)$$

where $f$ is in megahertz, $L$ in microhenries, and $C$ in picofarads, shows that capacitance is the inverse ***square*** of frequency. If you only think in terms of band limits, then the capacitance ***ratio*** is equal to the ***square*** of the frequency *ratio*.

Although the actual value of the L and C components is important in determining the impedance of the circuit, only the ratio of the variable capacitor will determine the ratio of the frequency range. Since it is usual practice to use variable capacitors in vfo design (as opposed to variable inductors), you can assume that inductance is only fixed or trimmable. The calculations will also give total circuit capacitance for determining circuit inductance and impedance. Following are several of the constants which are used in the bandspreading calculations:

$$D = \text{Desired capacitance ratio} = (f_{max}/f_{min})^2$$

where $f_{max}$ and $f_{min}$ are the band limits desired

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$V$ = Variable capacitor's capacitance ratio, maximum to minimum
$C_v$ = Variable capacitor, maximum value
$C_p$ = Capacitance in parallel with $C_v$
$C_s$ = Capacitance in series with $C_v$
$BW$ = Tuning bandwidth, a function of $D$
$d$ = $\sqrt{D}$, used in the bandwidth equation

A few other specific terms will be defined later.

The desired capacitance ratio, $D$, may vary slightly from its original value. This depends on the tolerances of $C_p$ or $C_s$, and whether or not trimmer capacitors are used. The final value of $D$ must be larger than the original $D$ if the desired frequency band is to be covered.

The tuning bandwidth may be checked by the following equation:

$$BW = (d - 1)\sqrt{\frac{f_{max} f_{min}}{d}} \tag{2}$$

Remember that $d$ is the square-root of the final capacitance ratio, $D$. **Eq. 2** assumes that the resulting tuning band has the same center frequency as the desired band. $BW$ has the same scale as $f$, so you can use either megahertz or kilohertz. Usage is shown in the examples which follow.

## parallel capacitor

$$C_p = \frac{C_v(V - D)}{V(D - 1)} \tag{3}$$

$$D = \frac{V(C_v + C_p)}{VC_p + C_v} \tag{4}$$

Total capacitance, $C_t$, at the lowest frequency, is $C_v$ + $C_p$ (see fig. 1). For example, a 40 - 360 pF variable

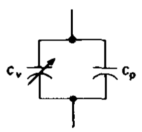


**fig. 1. Parallel-capacitor bandspreading circuit. Total capacitance, $C_t = C_v + C_p$.**

capacitor is to be used in a vfo to cover the entire 80-meter band. Constants are:

$$D = (4.0/3.5)^2 = (1.143)^2 = 1.306$$

$$V = 360/40 = 9$$

$$C_v = 360\ pF$$

Plugging in the constants,

$$C_p = \frac{360(9 - 1.306)}{9(1.306 - 1)} = \frac{2769.8}{2.7549} = 1005.41$$

A 1000 pF fixed capacitor could be used if the tolerance was 0.5 per cent and no stray capacitance was present in the circuit. A better choice would be to use a fixed capacitor and a trimmer.

Using 5% 680 pF and 220 pF units in parallel, the total capacitance, at maximum tolerance is 945 pF (requiring about 60 pF minimum trimmer and stray capacitance). Total parallel capacitance at minimum tolerance is 855 pF (requiring about 150 pF trim and stray). Without any trimmer, and assuming a stray capacitance of 5 pF, the nominal value of the parallel

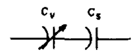


**fig. 2. Single series capacitor should be used with care in bandspread circuits because this arrangement does not allow for any stray capacitance.**

capacitance, $C_p$, is 905 pF. Tuning bandwidth is slightly wider and the new value of $D$ is

$$D = \frac{9(360 + 905)}{(9 \cdot 905) + 360} = \frac{11385}{8505} = 1.339$$

The square-root of $D$ is 1.16 = $d$. Using the tuning bandwidth equation

$$BW = (1.157 - 1.0)\sqrt{\frac{14.0}{1.157}}$$

$$= 0.157 \cdot 3.479 = 0.546\ MHz$$

The resulting bandwidth is less than 10 per cent wider than desired, using the nominal value of $C_p$.

The maximum tolerance of $C_p$ yields $d$ = 1.150 and $BW$ = 0.524 MHz, while the minimum yields $d$ = 1.164 and $BW$ = 0.570 MHz. These are worst-case conditions and the difference in tuning bandwidths is only 45.28 kHz. Fixed values of $C_p$ could be used with a hand-calibrated dial for the variable capacitor, $C_v$.

## series capacitor

A single series capacitor should be used with care because this arrangement does not allow for stray capacitance across both $C_v$ and $C_s$. That will be discussed later. Equations for $C_s$ (fig. 2) are:

$$C_s = \frac{C_v(D - 1)}{V - D} \tag{5}$$

$$D = \frac{VC_s + C_v}{C_s + C_v} \tag{6}$$

If your calculator does not have a reciprocal function, the combined capacitance of the circuit, $C_c$, can be calculated from the following:

$$C_c = \frac{C_v(D - 1)}{V - 1} \tag{7}$$

To illustrate, let's use the same 40-360 pF variable capacitor and 80-meter band as before:

$$C_s = \frac{360(1.306 - 1.0)}{9.0 - 1.306} = \frac{110.16}{7.694} = 14.32\ pF$$

A 15 pF fixed capacitor could be used but the tolerance

would have to be better than 5 per cent because a low limit value of a 5% capacitor would be 14.3 pF, which is too low. However, parallel connected 5% 10 pF and 5.6 pF units will work with worst-case values of 14.8 and 16.4 pF.

To make certain that the entire band is covered, $C_s$ should be slightly larger than the calculated value. In the parallel situation, $C_p$ should be slightly lower than calculated. Remember these conditions when working with the series-parallel configurations which follow.

The new D value for the nominal 15.6 pF series capacitor is

$$D = \frac{(9.0 \cdot 15.6) + 360}{15.6 + 360} = \frac{500.4}{375.6} = 1.33$$

Tuning bandwidth will be 537.17 kHz for the nominal value of $C_s$. The maximum combined capacitance is then

$$C_c = \frac{360(1.33 - 1.0)}{9.0 - 1.0} = \frac{119.62}{8} = 14.95\ pF$$

## series-parallel configuration

Fig. 3 shows two ways of arranging $C_v$ and $C_s$ in a bandspreading circuit. The arrangement of fig. 3A is preferred because strays can be included with the parallel capacitor. The version in fig. 3B is useful where the series capacitance is part of a bypass or a feedback divider for a Clapp oscillator. Each version uses the same basic equations already discussed, but with different component notations to prevent any confusion.

Calculation can be done in two ways: either break up the circuit into the basic arrangements and use the equations already given; or solve directly with the equations which follow. In practice the latter process is a bit quicker.

In the series-parallel circuit of fig. 1A the component notations are as follows:

$C_{pa}$ = Parallel capacitance
$C_{ta}$ = Total maximum capacitance
$C_{cs}$ = Maximum capacitance of series part only
$A$ = Intermediate capacitance ratio, series part only

All other notations are as before. The $A$ value may be anything at the beginning but *must be between* $V$ and $D$ for the circuit to work. Since $C_s$ is usually a fixed capacitor

$$A = \frac{V\ C_s + C_v}{C_s + C_v} \qquad (8)$$

Then

$$C_{pa} = \frac{C_v\ (A - 1)(A - D)}{A(D - 1)(V - 1)} \qquad (9)$$

$$C_{ta} = \frac{DC_v(A - 1)^2}{A(D - 1)(V - 1)} \qquad (10)$$

Note that the denominators of $C_{pa}$ and $C_{ta}$ are the same. Storing the denominator in the calculator's memory allows a fairly quick calculation of both capacitance values.

In some cases the resonant impedance dictates a certain value of $C_{ta}$, and $C_v$ is the only other known value. The value of $A$ can be found by letting $C_j = 2DC_v$

$$C_k = C_j + C_{ta}\ (D - 1)(V - 1)$$

then

$$A = \frac{C_k + \sqrt{C_k^2 - C_j^2}}{C_j} \qquad (11)$$

and

$$C_s = \frac{C_v(A - 1)}{V - 1} \qquad (12)$$

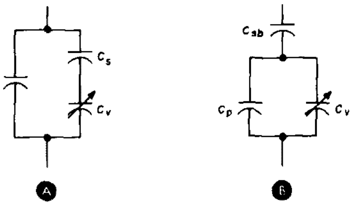


**fig. 3. Two basic series-parallel capacitor configurations used in bandspreading circuits. The circuit in (A) is preferred because strays can be included with the parallel capacitor $C_{pa}$. Arrangement in (B) is often seen in the feedback divider of a Clapp oscillator.**

To illustrate, an oscillator is required to cover 5.0 to 5.5 MHz with a 40 to 360 pF variable capacitor. The oscillator tank-circuit requires about 450 pF total at the low-frequency end.

$D = (5.5/5.0)^2 = 1.21$
$V = (360/40) = 9$
$C_v = 360$ pF
$C_{ta} = 450$ pF (approximately)

Solving for the approximate value of $A$

$C_j = 2 \times 1.21 \times 360 = 871.2$
$C_k = 871.2 + (450[1.21 - 1]\ [9 - 1]) = 1627.2$
$C_k^2 - C_j^2 = 2647780 - 758989 = 1888791$

$$A = \frac{1627.2 + 1374.3}{871.2} = 3.445$$

$$C_s = \frac{360(3.445 - 1)}{9 - 3.445} = \frac{880.3}{5.555} = 158.5\ \text{pF}$$

Therefore, 150 pF is the closest standard value. Using the first equation for $A$

$$A = \frac{(9 \cdot 150) + 360}{150 + 360} = \frac{1710}{510} = 3.353$$

$$C_{pa} = \frac{360(3.353 - 1)(3.353 - 1.21)}{3.353(1.21 - 1)(9 - 1)}$$

$$= \frac{1815.2}{5.633} = 322.25\ \text{pF}$$

$(A - 1)^2 = 5.536$

$$C_{ta} = \frac{1.21 \cdot 360 \cdot 5.536}{5.633} = \frac{2411.5}{5.633}$$

$= 428.13$ pF

This is very close to the required total tank capacitance. If $C_s$ is a fixed 5% unit, will a fixed value of $C_p$ with the same tolerance work? The lowest value of $C_p$ occurs with $C_s$ low:

$$C_s\ low = 0.95 \cdot 150 = 142.5 \text{ pF}$$

$$A = \frac{(9 \cdot 142.5) + 360}{142.5 + 360} = 3.269$$

$$C_{pa} = 306.18 \text{ pF}$$

$$C_{ta} = 408.27 \text{ pF}$$

The required value of $C_{pa}$ has decreased nearly 5 per cent from the nominal value of 322.25 pF. In order to get the variable's dial to cover as much of the band as possible, a fixed 270 pF capacitor with a parallel trimmer would be a better choice. The trimmer capacitor should have sufficient range to compensate for any tolerances in the 270 pF capacitor as well.

Where fixed capacitors are used for both $C_s$ and $C_{pa}$, *the narrowest* bandwidth will occur when $C_s$ is low and $C_{pa}$ is high. The widest bandwidth occurs with $C_s$ high and $C_{pa}$ low. To find $D$ for the bandwidth formula use the following relationship:

$$D = \frac{AC_{ta}}{C_{ta} + C_{pa}(A-1)} \tag{13}$$

In the series-parallel configuration of **fig. 3B** the component notations are:

- $C_{sb}$ = Series capacitance
- $C_{tb}$ = Total maximum capacitance
- $C_{pv}$ = Maximum capacitance of parallel part only
- $B$ = Intermediate capacitance ratio, parallel part only

As with the previous configuration, $B$ must have a value between $V$ and $D$. Note that there is no compensation for stray capacitance across the total; this arrangement is fine for circuits such as a Clapp oscillator where $C_{sb}$ is the total of the feedback divider network. Then

$$B = \frac{V(C_v + C_p)}{C_v + VC_p} \tag{14}$$

and

$$C_{sb} = \frac{B\,C_v(V-1)(D-1)}{V(B-1)(B-D)} \tag{15}$$

$$C_{tb} = \frac{B\,C_v(V-1)(D-1)}{V(B-1)^2} \tag{16}$$

Most circuits will have a fixed value of $C_{sb}$ so will require a direct solution for the value of $C_p$. Using temporaries

$$C_q = 4\,V C_v \left[ \frac{C_{sb}(V-D)}{D-1} - C_v \right] \tag{17}$$

$$C_r = C_v(V+1) + VC_{sb} \tag{18}$$

then

$$C_p = \frac{\sqrt{C_q + C_r^2} - C_r}{2V} \tag{19}$$

Taking the same oscillator example as before, with the same tuning range of 5.0 - 5.5 MHz and 40 - 360 pF variable, but in a Clapp configuration with two parallel 1000 pF feedback divider capacitors:

$$C_{sb} = 500 \text{ pF}$$

$$D = (5.5/5.0)^2 = 1.21$$

$$V = 360/40 = 9$$

$$C_v = 360 \text{ pF}$$

$C_p$ is found via the temporaries:

$$C_q = 4 \cdot 9 \cdot 360 \left[ \frac{500(9.0 - 1.21)}{1.21 - 1} - 360 \right] = 2.357 \times 10^8$$

$$C_r = 360(9+1) + (9 \cdot 500) = 8100$$

$$C_q + C_r^2 = (2.357 \times 10^8) + (6.561 \times 10^7) = 3.013 \times 10^8$$

$$C_p = \frac{17358.3 - 8100}{2 \cdot 9} = 514.35 \text{ pF}$$

The inductor must resonate with the total capacitance so the conventional series calculation with $C_{pv}$ = 874.35 pF (360 pF + 514.35 pF) would yield $C_{tb}$ = 318.10 (874.35 pF in series with 500 pF). Solving for $B$

$$B = \frac{9(360 + 514.35)}{360 + (9 \cdot 514.35)} = 1.577$$

$$C_{tb} = \frac{1.577 \cdot 360(9-1)(1.21-1)}{9(1.577 - 1.0)^2} = 318.31 \text{ pF}$$

The two $C_{tb}$ values differ by less than 0.1 per cent. Checking the worst-case tolerances of $C_{sb}$ will give $C_p$ = 501.55 pF and $C_{tb}$ = 306.19 pF with $C_{sb}$ at the −5% tolerance limit and $C_p$ = 526.70 pF and $C_{tb}$ = 329.76 pF at the +5% limit. Since stray capacitance in a Clapp oscillator tank circuit is primarily across capacitor $C_{sb}$, it can usually be neglected.

## things aren't always what they seem

**Fig. 4** illustrates what really occurs in a resonant circuit when the inductor exhibits appreciable distributed capacitance. This is more apt to be the case at lower frequencies where the inductance is large.

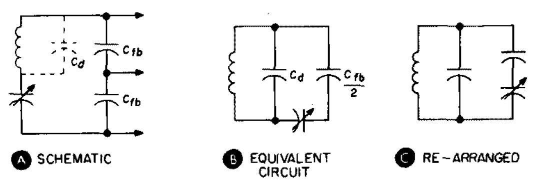


**fig. 4. Effect of distributed capacitance which appears across all practical inductors. In the circuit schematic (A), the distributed capacitance is shown as the stray, $C_d$, across the inductor. This is equivalent to the arrangement in (B) where the feedback capacitors have also been combined into one unit. The re-arranged circuit (C) indicates that the circuit of fig. 3A should be used when making bandspread calculations.**

Fig. 4A shows a Clapp oscillator tank as it would be drawn in a schematic. $C_d$ is the distributed capacitance across the inductor. In fig. 4B the distributed capacitance is shown and the feedback network is reduced to a single capacitor, $C_{fb}/2$. The equivalent circuit in fig. 4C

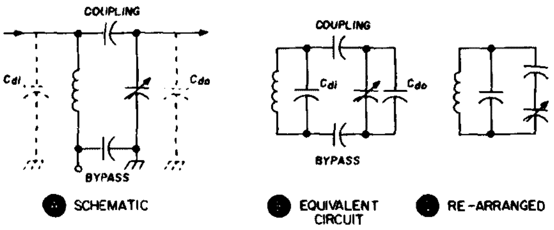


**fig. 5. Evolution of a tuned interstage circuit from the schematic (A), to the equivalent circuit (B), to the final configuration in (C) which indicates that the arrangement of fig. 3A should be used when making bandspread calculations.**

indicates that the series-parallel circuit of fig. 3A should be used when determining capacitor values. Distributed capacitance becomes $C_{pa}$ and the feedback divider becomes $C_s$.

Fig. 5 shows the transformation of a tuned interstage circuit with a bypass and coupling capacitor plus strays. Redrawn in fig. 5B, stray capacitance at the output, $C_{do}$, appears across the variable, and input strays, $C_{di}$, are across the inductor. In fig. 5C the variable capacitance is modified by the parallel output stray capacitance, and the bypass and coupling capacitors are combined in series with the variable. This fits the configuration of fig. 3A.

If you assume that the bypass, coupling, and stray capacitance will not affect tuning, you are wrong. Suppose a 10-50 pF variable capacitor was originally chosen along with 1000 pF coupling and bypass capacitors; strays are 5 pF on each side. Instead of the original capacitance ratio of 5, $D$ is now 2.788, and the tuning bandwidth is only 74.7 per cent!

## phase-locked-loop applications

An L-C oscillator with a varactor diode as the tuning element is often used as a voltage-controlled oscillator (vco) in a phase-locked loop because it has better noise characteristics than the emitter-coupled multivibrator type found in many *PLL ICs*.

The series-parallel circuit of fig. 3A applies itself well in this circuit. The varactor or VVC (voltage-variable capacitor) diode is $C_v$, $C_s$ can be a fixed value, and $C_{pa}$ can be trimmed or be fixed with the inductor trimmed. One caution: To keep the varactor under control of its bias voltage, peak rf voltage across the diode should not exceed the bias voltage. Also, the combined rf peak and bias cannot exceed the breakdown ratings of the diode. The smallest possible value of $C_s$ will prevent these conditions because the rf tank voltage across the diode is reduced by the fraction $C_s \,/\, (C_s + C_v)$. Remember, too, that bias voltage varies inversely to capacitance.

Most inductors have a positive temperature coefficient and most fixed capacitors have zero or negative temperature coefficients. A breadboard oscillator built with fixed, NPO type capacitors (zero temperature coefficient) will help to determine the amount of correction required.

The breadboard should be attached to a fairly large metal or ceramic mass to keep circuit temperatures from changing too rapidly. A casserole dish with a cover is handy for the thermal mass. The freezer compartment of a refrigerator can serve as a "cold soak" chamber. An all-glass thermometer is good enough if the bulb end is mounted near the circuit. A transparent dish cover allows you to see everything; masking tape can be used to seal the openings for the power and output wires. A frequency counter or calibrated receiver is essential for frequency-drift measurements.

Decreasing frequency with increasing temperature means that negative temperature coefficient capacitors must be used to compensate the circuit. The following formula works in this case and where fixed parallel capacitors are used

$$C_{vt} = \frac{C_{tot}\,(D_t - 1)\,10^6}{D_t\,\Delta T_{tc}} \tag{20}$$

Where:

$C_{tot}$ = total capacitance across the inductor

$C_{vt}$ = that part of $C_{tot}$ which requires temperature compensation

$D_t$ = square of the frequency change, greater than one

$\Delta T$ = temperature range of measurement, degrees Celsius

$tc$ = negative temperature coefficient expressed as parts per million (ppm) per degree C

For example, suppose an oscillator circuit has $C_{tot}$ = 1000 pF, made up largely with fixed NPO values, a 40°C temperature range, and N750 type ceramics are to be used for compensation. Frequency drift is 5.000 to 5.063 MHz. Therefore, the highest to lowest frequency ratio is 1.0126 making $D_t$ = 1.02536.

$$C_{vt} = \frac{1000(1.02522 - 1.0)\,10^6}{1.02536 \cdot 40 \cdot 750}$$

$$= \frac{2.536 \times 10^7}{30760.8} = 824.4 \text{ pF}$$

This is within about 0.5 per cent of the standard value of 820 pF, which would provide acceptable results for most applications. In some cases you may find that, with some trial and error calculations, you can arrive at a combination of fixed compensating capacitors which will do the job.

The casserole dish method will take about three hours for frequency measurements to stabilize, with about the same amount of time required for the cold soak. The dish can be elevated in temperature with a couple of lamps heating the sides of the dish. Turn the lamps off at

least ten minutes before making a frequency measurement to allow the "chamber" to stabilize.

For best results, the dish should have opaque sides. Pyrex dishes work all right if a heating pad is substituted for the lamps. The difference in heaters is probably due to infrared heating of the breadboard directly through the transparent Pyrex. It is also possible that the thermometer can be heated more by direct lighting, thus giving a false reading.

## measurement of unknown values

A bridge or Q-meter is the obvious choice for measuring unknown capacitance but good accuracy can be achieved with simpler methods. A grid dipper and one standard capacitor will serve as a starter. A standard in the 100 to 200 pF range is acceptable for most work. Arco-ElMenco has a series of 1-percent standards which are an excellent choice.

The dipper method uses a test inductor that resonates with the standard capacitor. The unknown capacitor is then resonated with the test coil and the two frequencies are measured. The unknown capacitor differs from the standard by the square of the resonant frequency ratio. Overall accuracy depends both on the standard capacitor and on the method (and accuracy) of frequency measurement.

Most grid-dipper dials are too coarse for even 5 per cent accuracy. Coupling to the test coil varies and the amount the dipper frequency is "pulled" depends upon how closely the dipper is coupled to the tuned circuit. An all-band receiver can be used to tune in the dipper for better frequency accuracy. Masking tape can be used to hold both the dipper and test coil in position.

If there is metal on or just under your workbench, elevate the grid dipper and the test coil about an inch (2.5cm) on a piece of Masonite or plywood to avoid excess capacitance coupling to the metal. The dipper case should be grounded to the receiver chassis to prevent dipper detuning when changing the position of your hands.

## making your own standard capacitor

There are plenty of old 30-365 pF broadcast-band variables in junkboxes. A three-gang variable can be used as a two-range standard variable for 20-243 and 90-1095 pF. A simple schematic is shown in fig. 6.

The variable should be insulated, mounted in a metal box, and provided with a good dial. A vernier dial with a numeric scale is fine since the device won't be used daily. Heavy wire, at least number 18 (1.0mm) should be used and the dpdt switch of the slide type for minimum inductance.

Before starting construction, arrange to have access to an accurate bridge for calibration of the finished standard. A Q-meter may be needed for the low range. Q-meters have accurate capacitance and the substitution method is used for calibration. Substitution involves a test coil on the Q-meter inductor terminals with the variable standard connected to the capacitor terminals. The maximum capacitance on the low range is measured on a bridge, the Q-meter capacitance is set at minimum, and peaked with the frequency dial. Calibration is then made by holding frequency constant, moving the Q-meter capacitor to a higher value, then repeaking by lowering the variable standard. The variable's capacitance from maximum is equal to the Q-meter's capacitance difference from its minimum starting point.

## distributed capacitance and true inductance

Large inductors have distributed capacitance which is parallel with the inductor. In high-impedance resonant circuits, this may affect tuning. The distributed capacitance also affects true inductance measurements. It should be considered in air-wound coils greater than 50 $\mu$H or greater for slug-tuned coils. Both distributed capacitance and true inductance may be determined by using a grid dipper (or Q-meter) and applying the following formulas.

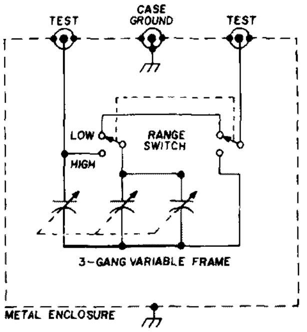


**fig. 6. Circuit for building your own standard capacitor using a three-section 365-pF variable. Capacitance is 20-243 pF in the low range, and 90-1095 pF in the high range.**

$C_2$ = maximum standard capacitance, pF, at lower frequency, $f_l$ (MHz)

$C_1$ = minimum standard capacitance, pF, at higher frequency, $f_h$ (MHz)

$C_d$ = distributed capacitance of inductor, pF

$L_t$ = true inductance, $\mu$H

then
$$C_d = \frac{C_2 - 4\,C_1}{3} \ pF \tag{21}$$

With $f_h$ twice the frequency of $f_1$

then
$$L_t = \frac{18997.7}{f_l^2\ (C_2 - C_1)} \ \mu H \tag{22}$$

## summary

Calculate carefully, allow for tolerances, and use short, thick connection leads to reduce stray inductance. The techniques given here will reduce many weekends of "cut and try" to a few hours with paper and pencil. *Increased accuracy is an added bonus.*

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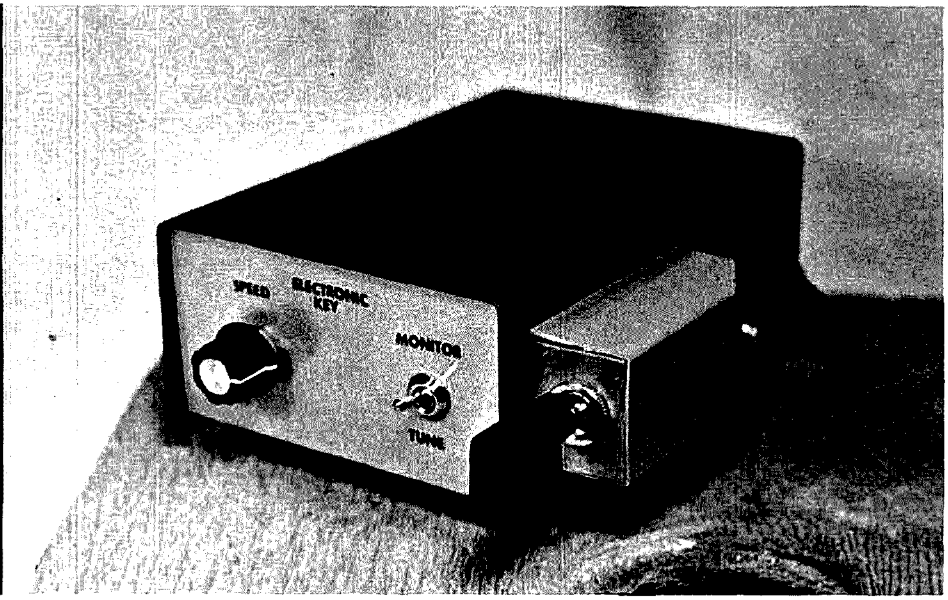


# ideas for a portable keyer paddle

Two designs for mechanical input to your electronic keyer — great for portable work

Often the need arises for a small, dependable key to complement a miniature portable electronic keyer for use with QRP transmitters. The keyer electronics can be small, lightweight, and dependable.[1]

Several keyers were designed and either mounted inside the QRP rigs or used outboard. However, the choice of keys to operate the keyers always left something to be desired. Two key designs were developed that are small, lightweight, and dependable. I've named these the *Dip-Key* and the *Lev-Key*. Each uses a different operating principle, but both function well. The *Lev-Key* is particularly rugged and survived a recent trip to a Pacific island while operating QRP portable.

## the dip-key

The *Dip-Key* schematic is shown in **fig. 1**. This circuit is basically a miniature touch-operated switch. The switch closure of the output transistor is sufficient to trigger the keyer inputs of cmos keyers. The heart of this design is a 14-pin dual in-line (DIP) wire-wrap socket, which is used for the two key levers. The DIP socket is mounted so that the pins extend out from a small piece of board. For a 14-pin socket, pins 1, 3, 5, 7, 8, 10, 12,

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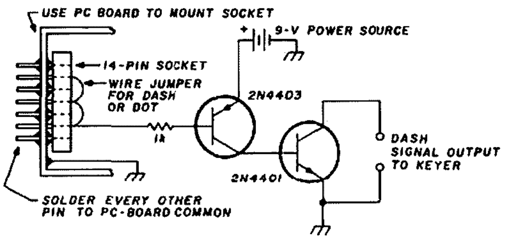
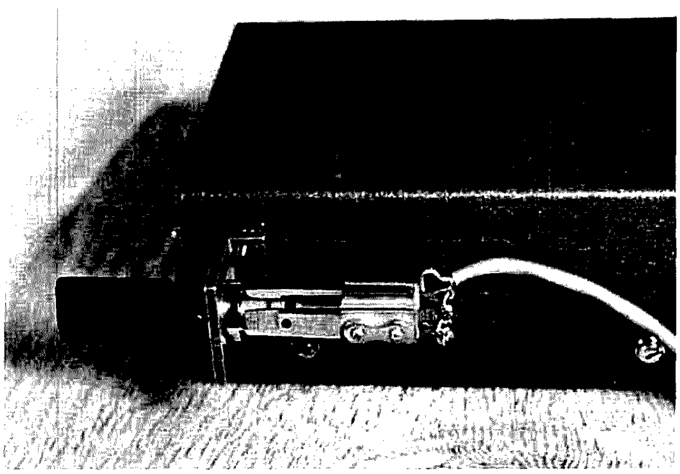



**fig. 1. One-half of the DIP-KEY circuit showing use of a 14-pin wire-wrap DIP socket for a paddle. Dash circuit is shown; dot circuit is the same design. Socket is mounted from the inside of a small enclosure with the pins protruding to the front. Battery is mounted with the circuit, or power can be taken from the electronic keyer.**

and 14 are soldered to the PC board ground. Pins 2, 4, and 6 are wired together to form the dash paddle, while pins 9, 11, and 13 are connected in parallel to form the dot paddle. Finger-skin resistance between any two adjacent pins creates a switch closure at the output of the respective dot or dash transistor. A small 9-volt battery was used to power the transistor switch, but power may be taken from the keyer power supply.

Features of this design are iambic twin-paddle operation, no moving parts, quiet operation, and extremely small size. The size can be as small as that of one 14-pin socket if the transistor and other components are mounted with the rest of the keyer electronics. The DIP socket is mounted from the inside of the mounting plate, with wire-wrap pins extending forward. Almost any transistor can be substituted for those shown. The only disadvantages to this design are the exposure of the small pins to the elements and the need for a fairly low skin resistance (less than 100k) for reliable switch operation. At times, with very dry fingers, it may not be possible to obtain a reliable switch closure. Normally, however, there seems to be low enough resistance, especially during contest operation!

**Side view showing LEV-KEY mounted on the QRP Accu-Keyer.**



## the lev-key

The *Lev-Key* evolved from the use of pushbuttons as the dot and dash levers for a keyer during portable operation. Pushbuttons are unreliable and difficult to use efficiently. I've discovered, however, that by using a standard switch known as a Switchcraft *Lev-R* switch, a dependable and reliable key can be constructed. The beauty of the *Lev-Key* is that mounting requires just one 0.5-inch (12.5mm) hole. The key is narrow and small. A three-position, spdt normally-open switch was chosen. The center position is *off*. Thus, by moving the switch lever one way or the other, a dot or dash switch closure is achieved. By judicious spacing of the switch contacts travel may be adjusted to individual preferences. This key is easily added to existing keyers or may be built into the rigs themselves.

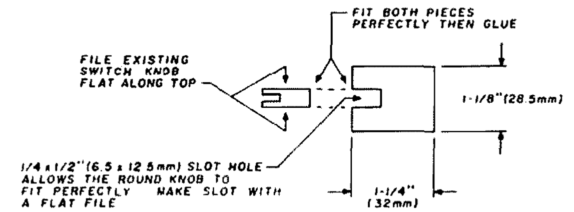



**fig. 2. Paddle extension for the LEV-KEY. The paddle, which is made from fiberglass PC-board material, is cemented to the round knob then painted black.**

The Switchcraft *Lev-R* switch is identified by the manufacturer's type number 3037 and consists of two sections. The switch lists in the Burstein-Applebee catalog for $3.92, but these switches may also be obtained from surplus equipments and junk boxes. If the *Lev-R* switch is not the spring-return type, a locking type will work by bending the leaf springs slightly so the leaf spring does not cause a locking action. The photos show the method used to mount the *Lev-Key* alongside the QRP Accu-Keyer.[1] Note the small paddle added to the original black plastic knob. Fig. 2 shows the dimensions for the added paddle. The features of this key are extreme ruggedness, small size, easy mounting, and low cost. The main disadvantage is single-paddle operation, which rules out iambic operation.

Although these designs may not be new, they give some ideas to those who wish to operate portable with minimum size and weight. These designs require only a very small investment for the fellow who would like to have a paddle for his keyer.

## reference


1. Gene Hinkle, WA5KPG, "The QRP Accu-Keyer," *QST*, January, 1976, page 24.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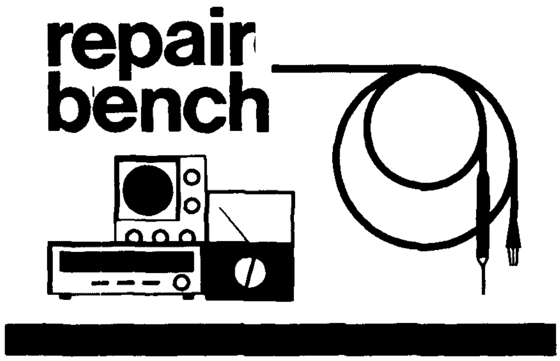


# Pat Shreve, W8GRG

## troubleshooting logic circuits

What is a logic circuit? In its broad sense, the term includes a wide range of devices, from simple analog units such as a volt-ohmmeter to the latest microprocessor. In this discussion I am going to take a narrower view, and will concentrate on troubleshooting methods applicable to simple digital logic circuits and devices using common ICs and discrete components which are likely to be found in amateur radio station equipment and accessories.

### understanding logic circuits

When the subject of digital logic and integrated circuits comes up, many amateurs back away with remarks such as, "I used to know quite a bit about vacuum tubes, and I've learned a little about transistors, but this IC stuff is all Greek to me." So let's not back away, but step up and see what sort of circuits are involved. Take for example the 7405 inverter, one of the basic units in the popular TTL (transistor-transistor-logic) line. An IC is commonly called a "chip", and the internal structure of one of the six inverters on this IC chip is shown in fig. 1. It consists of three transistors and three resistors, with a reverse-voltage protection diode placed between the input and ground. The transistors and resistors are connected in such a way that when the input is grounded Q1 is turned on and Q2 is turned off; which turns off Q3, and ungrounds the output. When the input is above ground the sequence is reversed and the output is grounded. Note the similarity between this circuit and the 7403 NAND gate shown in fig. 2. The only difference is that the 7403 has a second emitter on Q1 to provide a second input. Grounding either input turns Q1 on, so the output is *ungrounded* unless both inputs are *high*, that is, above ground.

Practically all TTL ICs are variations and combinations of these, or *similar*, gate and inverter circuits. If you can understand them, you can understand TTL logic. This is important, because the key to troubleshooting any logic circuit is understanding how it is supposed to work and what the various components are supposed to do.

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### troubleshooting tools

Contrary to many amateurs' impressions, you don't need a lot of expensive equipment to analyze and repair most of the logic circuits found in amateur stations, repeater control systems, autopatches, etc. It is true that if you are going to work on high-speed synchronous logic, time factors are important and a dual-trace, high-frequency oscilloscope is a big help, if not a necessity. Most of us are not going to be servicing a computer, however. The logic with which we are concerned either operates slowly enough or can be slowed down enough to permit checking the logic state of inputs and outputs with a simple voltmeter. It doesn't even have to be an expensive meter – one of the nice things about digital logic is that, when operating properly, inputs and outputs are either *high* (close to the 5-volt supply potential) or *low* (close to ground potential). Any in-between voltage is an indication of possible trouble.

At the occasional test point of a timer or counter reset where the logic is supposed to generate a brief pulse, a simple logic probe is useful. The one I use was described in *ham radio*, and the circuit is reproduced in fig. 3.[1] It can be built on a PC board or small strip of Vector board and enclosed in any convenient plastic tube. I used the body of a disposable 20cc plastic syringe discarded by a local laboratory. One other thing you will need is a manual or individual data sheets on the ICs with which you are working. These are obtainable from the manufacturer or distributor.

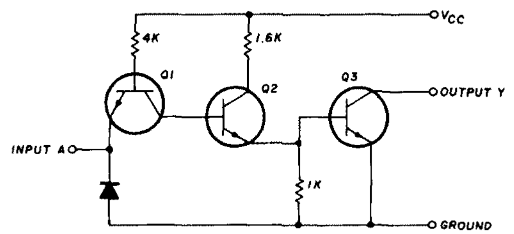



**fig. 1. Typical circuit of one of the six inverters in a SN7405 IC. Component values are nominal, not exact.**

### checking a simple circuit

Let's assume you have assembled your probe. Before you put it in the plastic tube, connect the red wire to the positive (+) terminal, and the black wire to the negative (−) terminal of a 5-volt source. If everything is okay, the LED should light when you apply +5 volts to the input with the switch in the positive position. In

either case it should stay lighted until you press the reset button.

Suppose the LED in the probe does not light at all. If you have any doubt about the LED, first test it and its 270-ohm series resistor by applying +5 volts to the resistor after disconnecting it from pins 3 and 4 of the IC. Check with a voltmeter to be sure there is +5 volts at the hot end of the 10k resistor, the 0.01 $\mu$F capacitor, and pin 14 of the IC as shown in the diagram. With an ohmmeter, check continuity to the black lead from pin 7 of the IC, the ground end of the capacitor, and the emitter of the transistor.

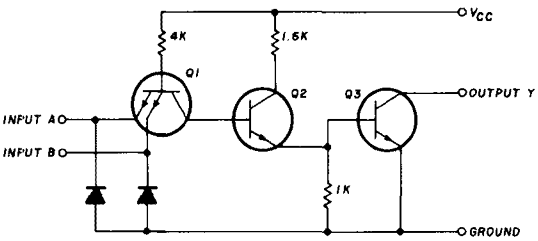



**fig. 2. Typical NAND gate; one of four in a SN7403 IC.**

Now connect the negative lead of the voltmeter to ground and the positive lead to the junction of the 10k resistor and the transistor collector. It should show zero potential when the input is grounded with the switch in the negative position, and also when the input is at +5 volts with the switch in the positive position. It should show +5 volts when the switch position or input polarity is reversed. If the circuit does not meet both tests, check the 10k and 4.7k resistors with an ohmmeter; if they are both okay, replace the transistor.

If the transistor, LED, and related components all check out, the trouble is in the IC. The simplest thing to do would be to replace it; but for practice, let's check it out. The 7400 gate is just like the 7403 illustrated in **fig. 2** except for an internal source of +5 volts to the output when it is in its *high* or logic 1 state. Thus pin 6 should be *low* when both pins 4 and 5 are *high*, and pin 6 should be *high* if either pin 4 or 5 is *low*. Pin 3 should be *low* when both 1 and 2 are *high*, and it should be *high* if either input is *low*. Connect the negative test lead of the voltmeter to ground and the positive lead to the junction of pins 6 and 2. Press the reset button. The meter should show more than +2.5 volts (typically 4 volts or more) when the button is pressed. It should stay at this level when the button is released until the input is grounded with the probe switch in the negative position or connected to V+ with the probe switch in the positive position. When either of these input conditions occurs the meter reading should drop to less than 0.5 volt. The LED should also light. If the circuit fails any of these tests, try a new IC.

## testing a more complex circuit

As an example of how to test a more elaborate system, consider the circuit shown in **fig. 4**. It is the access control device used on the Lake Erie Amateur Radio Association 16/76 repeater in Cleveland, Ohio, which I described in an earlier article.[2] It is a good circuit to illustrate troubleshooting procedures because it contains a timer, two counters, NAND and NOR gates, and some discrete components – all on one circuit board.

First, let's look at what the circuit is supposed to do when it is operating normally. If there is no outside interference on the input frequency, the control unit permits the repeater to run *open*, i.e., carrier access without an access tone. The receiver COR (carrier operated relay) pulse is shaped by the 7400 gates U5A and U5B, and keys the transmitter through U5C as long as the output of U5D is *high*. Note that turning the guard control switch *off* latches the output of U5D *high* and lets the repeater run open regardless of what is happening in the rest of the unit.

If the control switch is *on*, the output of U5D is controlled by U1. This is a 7490 decade counter which is really two counters: one that divides by two and *one* that divides by five. For either to count, pin 2 or pin 3 and, at the same time, pin 6 or pin 7 must be *low* (grounded). (Remember, TTL logic sees unconnected inputs as *high*). With proper input conditions, each cycle of the COR registers one count on the divide-by-five counter. Pin 9 goes *high* on the first count, pin 8 on the second, both 8 and 9 on the third, and pin 11 on the fourth. On the fourth count U1 latches with pins 11 and 12 *high* until the counter is reset. When latched with the control switch *on*, U1 prevents keying of the transmitter by the COR. Resetting is accomplished either by a signal with an access tone through one of the diode inputs, the discrete components, and the U2 NOR gates; or by the

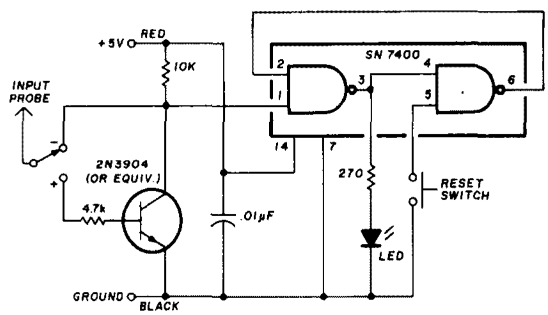



**fig. 3. Simple logic probe. Parts arrangement is not critical. See reference 1 for typical layout.**

timer circuit of U3 and U4, which also works through the gates.

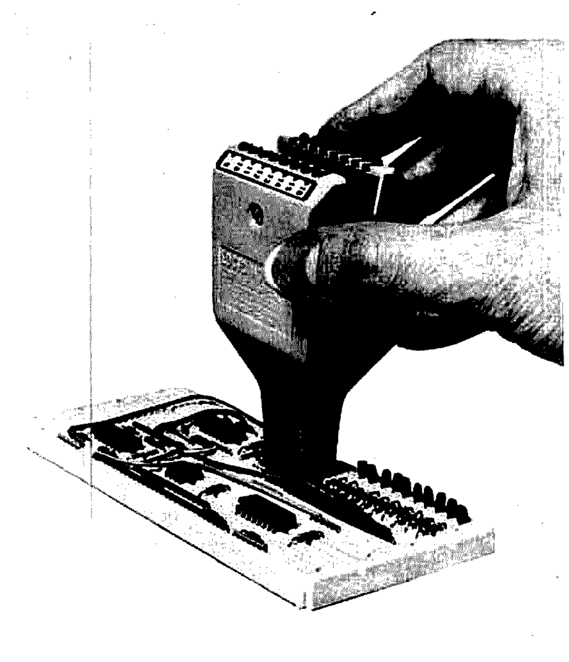
To illustrate the troubleshooting procedure, let's consider three possible outward signs of a malfunction: failure to guard the repeater by cutting off the keying circuit in the event of repeated interference, failure to respond to an access tone, and failure to revert to carrier access at the end of the interval for which the timer is set. The pulse-shaping and keying circuits of U5 are similar to the logic probe circuit analyzed earlier, so let's assume that those circuits are working as they should.

If the system does not go into the guard mode after being keyed by four successive signals without an access tone, U1 is probably not counting the way it should. This could be caused by a defect in the counter IC, U1, or by an improper signal from some other part of the circuit. Failure to reset could also be due to U1, or due to one of the gates or some other part of the circuit. Since everything comes together at U1, it is a good place to start looking for trouble.

## checking the counter

Before doing any testing it is a good idea to disconnect the inputs from the receiver so your checking won't be disrupted by someone keying the repeater. Check the supply voltages with a meter to be sure there is +5 and +12 volts at the power connections to the circuit board and the indicated pins 8, 9, 11, and 12 of U1. All should have less that 0.5 volts or should go to that level when pin 7 is grounded. If they don't, take a reading at pin 3 while pin 7 is grounded. U2B is connected as an inverter between pins 7 and 3 of U1, and pin 3 should be high when pin 7 is low. If it isn't, replace U2. If it is, and the voltage at any of the output pins is still above 0.5 volts, U1 is bad and should be replaced.

To check how U1 counts, you should start with all the outputs and pin 3 *low*, pin 7 *high*. Drive pin 1 *high* by momentarily grounding pin 1 of U5A. Then touch pin 1 of U1 with a grounded probe. Pin 9 should go *high*. Touch pin 1 again; pin 9 should go *low* and pin 8 should go *high*. Both should be *high* after the next pulse, and pin 11 should go *high* on the fourth pulse. When pin 11 goes *high*, so does pin 6. With both pins 6 and 7 *high* the counter should latch, with highs at pins 11 and 12, and *lows* at pins 8 and 9. Additional pulses applied to pin 1 should have no effect on the outputs. If it doesn't latch, check pin 7; if pin 7 is high and pin 12 is not, replace U1.



**Continental Specialties Logic Monitor LM-1 facilitates checking digital ICs and associated circuits. Priced at $84.95 from Continental Specialties Corporation, 44 Kendall Street, Box 1942, New Haven, Connecticut 06509.**

Pin 10 on U2C controls pin 7 on U1 and pins 5 and 6 on U2B. It should be high unless either pin 8 or pin 9 (the U2C inputs) is *high*. Pin 9 should be *low* at all times except for a momentary pulse when the timer resets. If pin 9 shows a continuous *high* there is trouble in the timer. If pin 9 is low, check the voltage at pin 10 when you ground pin 8. Pin 10 should go *high* when pin 8 is *low*, and vice versa. If it doesn't, replace U2.

Pin 8 should be *low* except when a signal with an access tone generates a *low* at input CR1 or a *high* at CR2. If it is not, there is trouble with U2D or in the discrete component circuitry. Check the voltage at pin 13 of U2 when you ground pin 11 or 12. If pin 13 doesn't go *high*, U2 is bad. If pin 13 stays *high* all the time, check pins 11 and 12. If pins 11 and 12 are also *high* replace U2. If pins 11 and 12 stay *low* there is trouble in the transistor circuit.

## checking discrete components

The reset circuit composed of CR1, CR2, Q1, Q2, and the associated resistors is intended to reset U1 by grounding pins 11 and 12 of U2 when a ground pulse is sensed at CR1 or +12 volts at CR2. With neither of these signals present, the voltage at the collector of Q2 should be approximately +5 volts. If it isn't, try grounding the base of Q2. If the collector voltage goes to +5 volts, Q2 and the 22k and 33k resistors are okay. If it doesn't, check each resistor. If they are the correct values, replace Q2.

With the input to CR1 ungrounded, Q1 should not conduct. Assuming the 12-volt supply (V+) is approximately right, the voltage at the junction of CR1 and the three resistors should be about 11 volts, and the voltage at the emitter of Q1 should be about 7.6 volts. There should be practically no voltage at the collector of Q1. If the first two voltages are about right, and there is significant voltage at the collector, Q1 is probably shorted and must be replaced.

The foregoing checks should have identified and corrected the reason for any failure of the control unit to guard the repeater, and should have identified most of the possible causes of control unit failure to respond to an access tone. If a signal with the correct tone still fails to open the guard circuit, apply +12 volts to the input of CR2. The voltage at the collector of Q2 should drop almost to zero. If it does not, CR2 could be failing to conduct, the 1k resistor could be shorted or there could be an open 4.7k resistor in the base circuit of Q2. These components can be checked individually.

If +12 volts at CR2 opens the guard circuit, but a ground at CR1 does not, try grounding the junction between CR1 and the 2.2k resistor. If this opens the guard circuit, replace CR1. If it does not, ground the base of Q1 through an external 2.2k resistor. If this

opens the guard circuit, replace the 2.2k resistor on the board. If it doesn't, Q1 is probably open and should be replaced.

### checking the timer

As previously mentioned, pin 9 of U2 is supposed to be *low* except for a brief pulse every 15 minutes when the timer resets. You can detect this pulse with the logic probe if you want to wait for it, but there are faster checks. Like U1, U3 is a 7490 decade counter. For it to count the pulses at its pin 14 input, it must have a *low* at pin 6 or 7 and at pin 2 or 3. Pins 6 and 7 are grounded, but pin 2 should be checked. If it is *high* go back and check U2 to see what is keeping it that way.

With pin 2 *low*, you can tap the junction between pin 14 of U3 and the 2.2k resistor with a grounded probe, and check the outputs of U3 with the logic probe or a voltmeter. Pin 12 should change from *low* to *high* or *high* to *low* every time you pulse pin 14. Pin 9 should change on every other pulse, pin 8 half that often, and pin 11 should go *high* on the eighth count. This last output can best be checked with the logic probe, as a *high* at pin 11 drives pin 10 of U2 *low* and pin 4 *high*. The resulting *high* at pin 2 of U3 will reset all its outputs, including pin 11, to the *low* state too fast for a meter to detect the momentary *high* at pin 11.

When you are satisfied with the operation of U3, move on to U4. Using an 8038 as a timer is not an economical way to do the job unless, as I did, you just happen to have one around. An NE555 is less expensive. You can connect your meter to the positive side of the 100 μF capacitor and watch the voltage rise and fall slowly as the timer cycles. Every time it changes direction the output at pin 9 should change state from *low* to *high* or back to *low*. If you don't see this action, test the 100 μF capacitor and the 47k and 100 ohm resistors. If they are all good, replace the IC.

### conclusion

If you have followed and understood each step of the testing procedure outlined here, you should have a pretty good idea of how to troubleshoot most logic circuits, even without elaborate test equipment. To summarize:

1. Review the data sheets on the ICs involved to be sure you know how each one should perform and how they affect each other.
2. Start at a point where you recognize a malfunction (usually an output), and work back to the source, step by step.
3. If you come to a point where the trail back to the input branches, temporarily disregard one branch and check out the other.
4. When you are satisfied that each component, or group of components, is responding to signals the way it should, mark it okay and go on to the next in line. When you finish a branch, go back and pick up the next one.

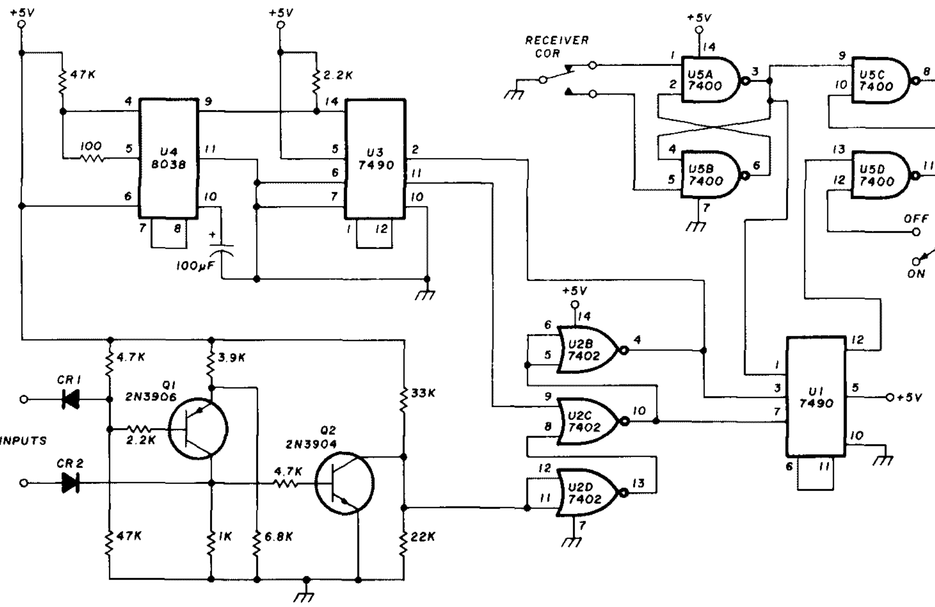



**fig. 4. Circuit diagram of a logic system used for repeater access control; a good example of the type of logic circuit found in many amateur installations. IC U2A is not used in this circuit.**

### references


1. Bill Rossman, "Logic Test Probe," *ham radio*, February, 1973, page 56.
2. R. B. Shreve, W8GRG, "Automatically Controlled Access to Open Repeaters," *ham radio*, March, 1974, page 22.


**ham radio**

MICROPROCESSORS

# microcomputer interfacing: internal registers within the 8080 chip

In this and several subsequent columns, we will introduce you to additional details concerning the operations of an 8080A-based microcomputer that are controlled by software. It is the software instructions, or steps, that actually indicate to the microcomputer the tasks it must perform. Just as you may start the day with a list of things to be done and a sequence in which they should be accomplished, the microcomputer too, must be provided with a sequential list of program steps. This is called *software*.

In general, you may not be familiar with what each microcomputer instruction does within the microprocessor chip. This should not deter you, however, from using them all in all of your programs. Many of you are not familiar with the inner workings of an internal combustion engine, an automatic transmission, or a Xerox machine; you lack of knowledge does not prevent you from using them daily.

All of our programs are stored in fast-semiconductor memory, either read-write memory or programmable

**By Peter R. Rony, Jonathan Titus, and David G. Larsen, WB4HYJ**

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read-only memory (PROM).[1] The microprocessor chip *fetches* an instruction byte from the memory and then executes it within the chip. Each software instruction requires at least one fetch and execute action. Software instructions that contain several instruction bytes require two or three sequential fetch and execute

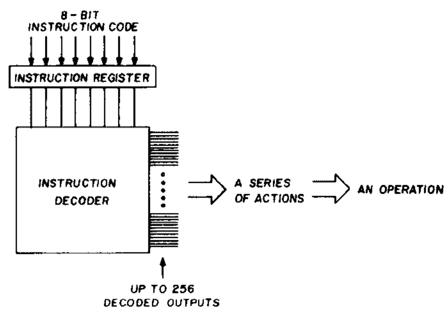


**fig. 1. The 8-bit instruction code is stored in the instruction register, where it is decoded into a series of actions that together comprise a microcomputer operation.**

actions. Again, exactly what is done within the microprocessor chip is not of interest, only the overall effect. All software is executed sequentially, one step after another, unless we purposely transfer control to instruction bytes located elsewhere in memory.

In order to conveniently handle the large number of software instructions in the 8080A microprocessor instruction set, it is customary to order them into several instruction groups. The Intel 8080 User's Manual subdivides the instruction set into five groups:

1. **Data Transfer Group.** Move data between registers or between memory and registers.
2. **Arithmetic Group.** Add, subtract, increment or decrement data in registers or in memory.
3. **Logical Group.** AND, OR, EXCLUSIVE-OR, compare, rotate or complement data in registers or in memory.
4. **Branch Group.** Conditional and unconditional jump instructions, subroutine call instructions and return instructions.
5. **Stack, I/O and Machine Control Group.** Includes I/O instructions, as well as instructions for maintaining the stack and internal control flags.

## registers

Before we can make sense of these instructions, we must know more about the internal architecture within the 8080 chip itself. We shall present such information in steps, since it can be overwhelming if tackled all at once.

Shown in fig. 1 is a schematic diagram that depicts the significant aspects of the internal architecture within an 8080 or 8080A* chip. Our emphasis in the diagram has been on accessible 8-bit and 16-bit *registers* that store information within the chip. You should exclude from consideration the Data Bus Buffer/Latch and the Address Buffer, which we show here to make the point that these are the internal circuits that interface the internal digital circuitry with the outside world, *i.e.*, the 8-bit *bidirectional data bus* and the 16-bit *address bus*. As you learn about 8080 microcomputer programs, you will be specially interested in the *accumulator*, *flags*, *program counter*, *stack pointer*, and *general purpose registers* B, C, D, E, H, and L.

We will not say much about the *instruction register*, since its function is automatic and we have little control over it. The function of the instruction register can be best understood with the aid of fig. 2 which indicates that it is the 8-bit register that temporarily stores the 8-bit instruction code for an instruction that is to be executed. Within the microprocessor chip, an instruction decoder converts the instruction code into a series of actions that together cause a microcomputer operation to occur. The individual actions are clocked by the $\phi_1$ and $\phi_2$ signals that are input at pins 22 and 15, respectively, on the 8080 chip.

The general purpose registers B, C, D, E, H, and L are used for many varied purposes, *e.g.*, the storage of an 8-bit constant, the storage of a 16-bit pointer address, the storage of an intermediate result in an addition or subtraction, etc. Each general purpose register is eight bits wide and can exchange data directly with the 8-bit external bidirectional data bus. Simple 8080 instructions permit you to transfer eight bits of data from one register to another, from a pair of registers to the program counter, from a register to the accumulator, and from a register to a memory location, and *vice versa*. You can use the contents of a register to perform addition, subtraction, AND, OR, EXCLUSIVE-OR, and compare operations with the contents of the accumulator. The contents or each register can be incremented or decremented. Register pairs, such as register H and register L, can be incremented or decremented as a 16-bit word.

The accumulator also acts as a general purpose register, but it has some special characteristics not possessed by the other six registers. The result of any arithmetic or logical operation is stored in the 8-bit accumulator. The I/O instructions IN and OUT transfer data only between the accumulator and external I/O devices. The contents of the accumulator can be transferred to any other general purpose register or to a memory location, and *vice versa*.

The five flags (zero, carry, parity, sign, and auxiliary carry) are flip-flops that indicate certain conditions have arisen during the course of an arithmetic or logical instruction. Such flags are used by the microcomputer in making decisions, *i.e.*, with conditional jump, call, and return instructions, in multiple-precision arithmetic operations, and in logical masking operations.

The program counter is a 16-bit register in the 8080 microprocessor chip that contains the address of the next instruction byte that must be executed in a program. You can load the program counter register either from a register pair or, more likely from two instruction bytes located sequentially in memory.

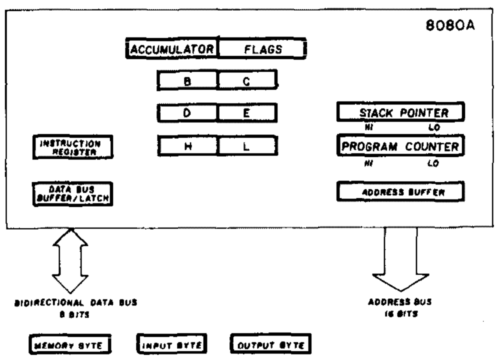


**fig. 2. The internal register architecture within an 8080A microprocessor chip. Temporary registers over which you have no direct control have been omitted.**

The stack is a region of memory that you allocate for the storage of temporary information, usually the con-

*The 8080A chip is an improved version of the 8080 chip. Both have identical instruction sets and pin outs. As one difference, the 8080A has improved fan-out capabilities.

tents of the internal registers within the 8080 chip. The stack pointer is a 16-bit register that contains the address of the last byte placed on the stack.

This summarizes our brief discussion of the architecture in a typical 8080 microprocessor chip (table 1). Keep in mind that there are seven 8-bit registers, two-

**table 1. Functions and registers within the 8080A microprocessor.**

| | |
|---|---|
| Accumulator | The register within a computer where the results of all arithmetic and logical operations are placed. |
| Address bus | A unidirectional bus over which digital information appears to identify either a particular memory location or a particular I/O device. |
| Bidirectional data bus | A data bus in which digital information can be transferred in either direction. With reference to an 8080A-based microcomputer, the bidirectional data path by which data is transferred between the microprocessor chip, memory, and input-output devices. |
| Fetch | In a computer, the collective actions of acquiring a memory address and then an instruction or data byte from memory. |
| Flag | A single flip-flop that indicates that a certain condition has arisen as, for example, during the course of an arithmetic or logical operation in a computer program. |
| General purpose registers | In the 8080 microprocessor chip, the 8-bit B, C, D, E, H, and L registers. |
| Program counter | In a computer, the register that contains the address of the next instruction byte that must be executed in a computer program. |
| Register | A short-term digital electronic storage circuit the capacity of which is usually one computer word or byte. |
| Software | The totality of programs and routines used to extend the capabilities of computers. Examples include compilers, assemblers, narrators, routines, and subroutines. |
| Stack | A region of memory that stores temporary information, usually the contents of the internal registers within a microprocessor chip. |
| Stack pointer | A register that contains the address of the last byte that has been placed on the stack in an 8080 microcomputer. |

16-bit registers, and five flags *the contents of which you can control using software*. Much of what an 8080 microcomputer does is to move 8-bytes from one location to another. This will become more apparent in subsequent columns.

## references

1. David G. Larsen, Peter R. Rony, Jonathan A. Titus, "The Structure of a Microcomputer," *American Laboratory*, October, 1975, page 111.

2. *Intel 8080 Microcomputer Systems User's Manual*, Intel Corporation, Santa Clara, California, September, 1975.

**ham radio**

# LC circuit calculations

## Table lookup for determining resonant frequency in terms of LC ratios

The resonant frequency of an LC circuit may be calculated from

$$f = \frac{1}{2\pi\sqrt{LC}} \quad (1)$$

where $L$ is in henries,
$C$ in farads, and
$f$ in hertz.

An approximate solution is sufficient in many cases, since errors may be present in the form of stray capacitance and self-inductance. Besides, the precise value of a coil or capacitor is seldom known anyway.

Presented here are some aids for solving the problem by using a simple table that allows you to determine resonant frequency in terms of LC products. Also included is a formula for finding inductance in microhenries of air-wound coils when coil diameter and number of turns are known.

An approximate solution may be obtained from an $L$, $C$, $f$, logarithmic graph. However, the $L$ and $C$ decades may list only values 1, 5, 10, 50, etc.; and the $f$ decades may be labeled only at 1, 2, 4, 7, 10, for example. Interpolating a log scale may be tricky, and any three-variable graph may be confusing.

**table 1. Basic data for determining LC products as a function of frequency.**

| $f$ | $L \times C$ | $f$ | $L \times C$ | $f$ | $L \times C$ | $f$ | $L \times C$ |
|---|---|---|---|---|---|---|---|
| 0.5 | 100,000 | 5 | 1000 | 50 | 10 | 500 | 0.1 |
| 0.6 | 70,000 | 6 | 700 | 60 | 7 | 600 | 0.07 |
| 0.8 | 40,000 | 8 | 400 | 80 | 4 | 800 | 0.04 |
| 1.0 | 25,000 | 10 | 250 | 100 | 2.5 | 1000 | 0.025 |
| 1.3 | 15,000 | 13 | 150 | 130 | 1.5 | 1300 | 0.015 |
| 1.7 | 9,000 | 17 | 90 | 170 | 0.9 | 1700 | 0.009 |
| 2.0 | 6,300 | 20 | 63 | 200 | 0.63 | 2000 | 0.0063 |
| 2.5 | 4,000 | 25 | 40 | 250 | 0.4 | 2500 | 0.004 |
| 3.0 | 2,800 | 30 | 28 | 300 | 0.28 | 3000 | 0.0028 |
| 4.0 | 1,600 | 40 | 16 | 400 | 0.16 | 4000 | 0.0016 |

### examples

A simpler method is illustrated in **table 1**. Here f is related to the product ($L \times C$). As an example, consider a circuit with 5 $\mu$H and 8 pF. The product (40) indicates a resonant frequency of 25 MHz. It's that easy. Try 15 $\mu$H combined with 20 pF. The product (300) gives 9 MHz. (Note: a *larger* product corresponds to a *lower* frequency).

**Table 1** works for MHz, $\mu$H, and pF but is not limited to these alone. It is also valid for kHz, mH, and nf, where $1\ nf = \mu F/1000$. Likewise it is valid for Hz, H, and $\mu$F (refer to **table 2**).

Here's another example: 500 $\mu$H x 250 pF. The product (125,000) shows about 0.45 MHz. Or you may simplify by writing 0.5 mH and 0.25 nf. The new product (0.125) corresponds to about 450 kHz. The table may be extended in either direction. You may multiply (or divide) $f \times 10$ when you multiply (or divide) the $L \times C$ by 100.

Let's use the table in reverse. We want a circuit to tune to 1000 Hz using an 88-mH coil. The LC product

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**table 2. Factors for using table with various units of frequency, capacitance, and inductance.**

| $f$ | $L \times C$ |
|---|---|
| Hz | H x $\mu$F |
| kHz | mH x m$\mu$F |
| MHz | $\mu$H x pF |

Note: $1\ m\mu F = \frac{\mu F}{1000}$

(table 1) is 0.025 and L is 0.088 H. The answer is 0.025/0.088 = 0.28 $\mu$F.

## determining inductances

The value of a capacitor is often known. If it is fixed, it probably has a color code or is labeled. A variable capacitor is known by the number of plates and size. A slug-tuned coil is specified as to maximum and minimum inductance values, but an air-wound single-layer coil is more mysterious. Its approximate inductance may be obtained from the formula

$$\mu H = \frac{D^2N^2}{18D + 40L} \tag{2}$$

where $N$ is number of turns,
$D$ is diameter, and
$L$ is length (both in inches).

For example, what is the inductance of 24 turns, 0.75 inch in diameter, and 1.5 inch long?

$$L = \frac{(0.75^2)\ (24^2)}{(18 \cdot 0.75) + (40 \cdot 1.5)} = \frac{324}{13.5 + 60}$$
$$\approx 4.4\ \mu H$$

**table 3. Some examples of inductances of air-wound coils based on eq. 2.**

| diameter, inch(mm) | turns, inch.(mm) | length, inch(mm) | $\mu$H |
|---|---|---|---|
| 0.5 (12.5) | 8 (204) | 2.0 (51) | 0.72 |
| 1.0 (25.5) | 8 (204) | 3.0 (77) | 4.20 |
| 2.0 (51) | 8 (204) | 5.0 (128) | 27.0 |

If the dimensions of the coil are given in millimeters, the approximate inductance in microhenries can be found from the following formula:

$$\mu H = \frac{D^2N^2}{460D + 1000L}$$

where $N$ is the number of turns,
$D$ is the diameter, and
$L$ is the length (both in millimeters).

For example, what is the inductance of 24 turns, 19mm in diameter, and 38 mm long?

$$L = \frac{(19^2)\ (24^2)}{(460 \cdot 19) + (1000 \cdot 38)} = \frac{207936}{8740 + 38000}$$
$$\approx 4.4\ \mu H$$

**ham radio**

# integrated-circuit tone generator

The new Mostek MK5085/6 IC can be applied to many different applications, most of which are usable by amateurs. The most interesting application though, is that it can be mated with a keyboard to form an inexpensive *Touch-Tone* generator. This chip will produce tones that are within 0.75% of the required frequency.

The circuit shown in fig. 1 requires a minimum of parts, all of which are readily available from local sources. A convenient feature of this IC is that it uses the standard 3.579545-MHz television color-burst crystal. These are low-cost crystals and even in this relatively remote area of northern Maine a call to a local television repair shop produced the crystal within an hour. This is much easier and cheaper than trying to purchase a 1-MHz crystal for other tone generators.

On the pin-out diagram (fig. 2), pin 15 is for invalid tone select. The pin would be grounded to provide dual tones only. This means that if a single tone condition is created (two keys in the same row or column selected) there would not be an output. Pin 10, mute switch, provides an output when a keyboard entry has been made. The transmitter switch output, pin 15, normally has a voltage present. During the times that keys are depressed, the voltage is removed. The mute switch can be used to key the PTT as shown in fig. 1. The capacitor

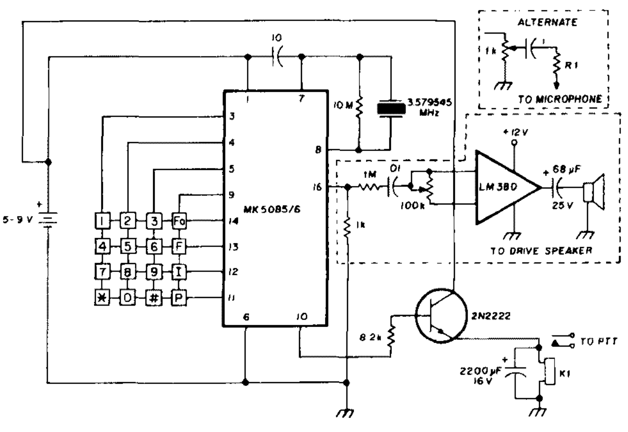



fig. 1. Schematic diagram of the complete Touch-Tone generator. The speaker can be eliminated and the output fed directly into the microphone input of a transmitter.

By Tim Ahrens, Post Office Box 895, Caswell AFS, Maine 04750

across the relay will keep it pulled in for approximately 2 seconds. Output (pin 16) is from the open emitter of a transistor. A load resistor, normally 1000 ohms, must be connected externally.

Depending upon the keyboard used, the correct IC can be selected from fig. 3. The MK5085/6-1 will operate with 3.5 to 10 volts applied while the MK5085/6-2 requires 4.5 to 10 volts dc. Current requirements are approximately 400 $\mu$A idling and 10 mA operating at 6 volts.

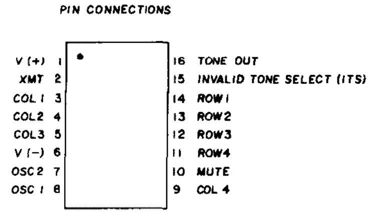



fig. 2. Pin-out of the MK5085/6. The ITS is grounded to eliminate any single tone outputs.

In my particular case, the internal portion of a 16 button pad was removed and a small printed circuit board substituted. This was not really necessary, but it provided more reliable tone generation than the original circuitry. To date, I believe this IC provides the average

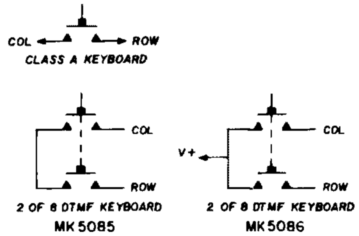



fig. 3. Keyboard connections to the MK5085/6.

ham the easiest way to generate correct tones for a minimum of expense.

## reference

1. Mostek data sheet, *MK5085N/MK5086N Integrated Tone Dialer*, Mostek Corporation, 1215 West Crosby Road, Carrollton, Texas 75006.

**ham radio**

# the ham notebook

## remote temperature sensor

Many times it is useful to know the temperature of an operational electronic device or component. An inexpensive temperature sensor which can be remotely installed inside a chassis or applied directly on a component to be checked, is a low cost silicon diode such as a 1N4004. The resistance of many diodes and other solid-state devices changes in a rather linear fashion vs temperature in the region from 32 to 212° Fahrenheit (0 - 100° C). A digital ohmmeter can be used to measure the "sensor" resistance (converted to temperature) after a simple calibration procedure has been completed and a chart similar to **fig. 1** is prepared. The actual measured resistances will vary from diode to diode, and from type to type, but it is *surprising how accurate these* simple sensors can be.

Calibration of a sensor is accomplished by applying crushed ice to a diode connected as shown in fig. 1 to establish 32°F (0°C) temperature equivalent resistance. The 212°F (100°C) point can be established by placing the sensor in boiling water. It is best to allow 3 minutes or longer at both calibration points for the device to stabilize. Distilled water gives the most accurate results but I have used tap water as well. If you wish to compare intermediate points by using an industrial thermometer, be sure to *stir* the water to equalize the temperature because there is usually a measurable temperature gradient, from top to bottom, in a container of water being heated. Draw a straight line between the values charted on a linear graph paper.

The curve in fig. 1 shows the small error (worst case 1.92%) of the sensor when compared to an expensive laboratory thermocouple meter.

**Max J. Fuchs, WA1NJG**

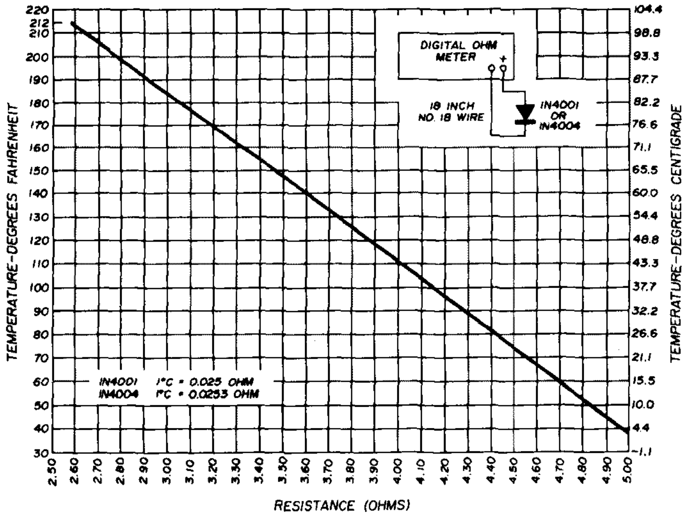



**fig. 1. Temperature vs resistance curve for low-cost 1N4004 silicon diode. Values for 1N4001 diodes are nearly the same. Worst case deviation from linear slope is 1.92%.**

## 75S-series crystal adapter

I recently acquired a Collins 75S-1 receiver. Navy MARS coverage was desired and I managed to locate a junk box 7081-kHz crystal that would temporarily allow me to cover the needed frequencies. It was a type HC-17/U which I was reluctant to alter to the HC-6/U package required by the receiver crystal board. There was also the problem of removing and replacing the crystals, their close spacing requiring needle-nosed pliers or similar device. I had no crystal/socket adapters on hand.

I solved the problem by removing the case and innards from a defunct HC-6/U crystal and soldering the tabs from a standard phenolic tube socket to the pins of the now empty HC-6/U crystal case. This construction alone has proven to be of sufficient strength although the assembly may be epoxied or incapsulated. With the adapter, HC-17/U crystals may be used (in fact, some FT-243s I tried also worked well), and are easily removed and/or exchanged.

**Paul K. Pagel, K1KXA**

## Collins 70E12 pto repair

After twenty-five years of faithful service my Collins 75A2 receiver quit. An extensive search uncovered the unhappy fact that its permeability-tuned vfo or pto had a shorted tubular, ceramic coupling capacitor.

Since factory service was out of the question in terms of time and money, I

decided to remove the pto myself and try to repair it. Here's what I did.

I used a short Philips screwdriver to remove the three screws holding the unit to its mounting plate in the receiver. In fact, I *made* a screwdriver to do this because I didn't happen to have one of the right size. You'll be luckier. I then snipped three wires and unsoldered the coax that goes to the 6BA7 second mixer tube. After I removed the pto cover, it was easy to replace the shorted capacitor with two 0.0047 $\mu$F silver mica capacitors.

I next replaced the pto unit — without its cover — in the receiver because I wanted to adjust its frequency calibration. Surprise! It was 8 kHz off when I re-installed it, but I was able to adjust it to within about 1 kHz. So far, so good, but now I had to remove the unit again to put its cover back on, and then re-install it a second time. Drat! Calibration was now 35 kHz off over the 1 MHz range. I removed it for the third time, and by now had become an expert in removing and replacing ptos.

Then I got a bright idea: Why not drill a small access hole in the pto can opposite the trimmer inductor, and also file a small screwdriver access notch in the pto mounting plate? I also oiled the lubricating washer with *Three-in-One* oil, and applied a dab of cold cream to the screw. I figured that if it's good for the face it shouldn't hurt the pto.

This time, after everything was back together, I made a small aluminum-wire tuning tool to reach the trimmer inductor through the hole I made in the pto can. Final calibration was less than 1 kHz from one end of the dial to the other. My 75A2 is now perking again better than ever, and I saved myself a whole lot of money in the bargain.

Oscar Tripancy, W6BIH

## safety circuit for pushbutton switches

A number of neat pushbutton switches has appeared on the surplus market, and these switches are being used for a variety of purposes around the ham shack. One common type is a bank of dpdt or 4pdt buttons ganged mechanically so that pushing one releases the others. This switch might be used, for example, to select one of a bank of crystals or to power up one of a group of circuits. Usual wiring ignores the normally-closed contact and applies the power (or ground) to all the commons in parallel. While the mechanical linkage releases the other buttons when one is pressed, it does not prevent simultaneously pressing (and locking) two buttons, thereby activating two separate circuits.

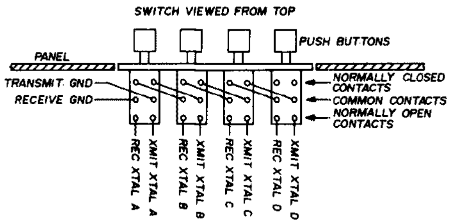



**fig. 1. Method of wiring multiple pushbutton switches so that two circuits cannot be connected at the same time, even though two buttons are pushed simultaneously.**

If activating two circuits is undesirable (as with a crystal selector), this possibility can be prevented by wiring the common of each pushbutton through the normally-closed contact of the previous switch so that the power (or ground) is carried as far as the first button depressed but not beyond (see fig. 1).

Richard Fleck, K3RFF
David McLanahan, WA1FHB

## ICOM-22A wiring change

Recently, while operating my IC-22A from the home station on 146.52 MHz simplex, I noticed the receiver had a slow recovery. Also, the in-line bridge indicated the emission of power at a low level. I further noted a low growling noise from the speaker at this time when I held my ear close to the speaker. Upon investigation of the circuit diagram and the wiring of the transceiver, I found what was an apparent wiring error.

PL1 and PL2 (the dial and S-meter lamps) are connected to the input side of the whine filter choke L35. It was also noted that another lead was connected to the same point on the terminal strip, from R3, the 100-ohm series resistor connected to PL3, the transmit indicator lamp. According to the wiring diagram, this lead should be connected to the *output* side of the whine filter (the junction of L35, C94, R102, and the common of S102).

After disconnecting the lead from the terminal strip and soldering it to the common of S102 (that side of the *high-off-low* switch which has an empty terminal) the problems disappeared. In addition, the small amount of alternator whine which was audible in the external speaker while mobile is now non-existent. Caution: Do not connect the wire from R3 to the empty terminal of S102, but to the *common of the switch* which is the *center terminal* on that side of the switch.

Paul K. Pagel, K1KXA

## low-cost copper antenna wire

I've found that copper antenna wire is either unavailable in the quantities that I need or is priced so high that I can't afford it; sometimes both.

Quite by accident an acceptable solution to the problem presented itself while I was in a local sporting-goods store looking for another item. A fisherman, who does a lot of lake-trout trawling, was buying stranded copper trawling wire in 300-foot (100m) lengths at what seemed to be a ridiculously low price. After he left, I questioned the proprietor about the availability of the line, and he answered, "No problem, I can get all I want."

Since then, I've become an avid antenna builder and use this material for all my wire antennas. The wire is a seven-strand, twisted copper wire that solders easily and has a diameter approximately equivalent to no. 16 AWG (1.3mm) wire. The only shortcoming I've found is that the wire will stretch under load over a period of time, sometimes as short as a month or so. Consequently I use it for inverted vees and other antennas where the wire doesn't have to bear all the load or strain of the connectors and coaxial cable. I've discovered, however, that I can easily trim or "prune" the wire to its orignal length after it has stretched for awhile. Stretching also appears to "cold-work" the wire slightly and is less pronounced after the initial deformation. Price? Well, the last 600-foot (200m) spool I bought cost only a bit over one cent per foot!

Jim Gray, W2EUQ


# comments

## memory keyer

Dear HR:

Since the popular WB9FHC memory keyer article was published[1] several modifications have been introduced for this relatively simple, low-cost circuit.[2,3,4] A significant disadvantage of the original circuit is the asynchronous clock, which allows sporadic keying anomalies. At low speeds a dot or dash may be lost if the paddle is hit and released between clock pulses. In addition, a "lost" dash may show up the next time the *dot* paddle is pressed. This phenomenon is the result of the master-slave flip-flop having an inherent memory at its J and K inputs. When the clock pulse is *high*, a brief *high* at the J input will be saved and clocked out when the clock goes low. While the clock is *low*, the J input will not accept a *high*. K3NEZ eliminated the unwanted dash by connecting the clock input of the dash flip-flop to the Q output of the dot flip-flop.[2] Thus, the dash flip-flop data inputs are disabled between characters by a low at the clock input.

At W6OCP, a different solution was found. Both disadvantages of the circuit were minimized by changing the duty cycle of the NE555V from a square wave to a series of short low-going pulses. Thus, the inputs to both the dot and dash flip-flops are enabled between clock pulses, and no characters are lost.

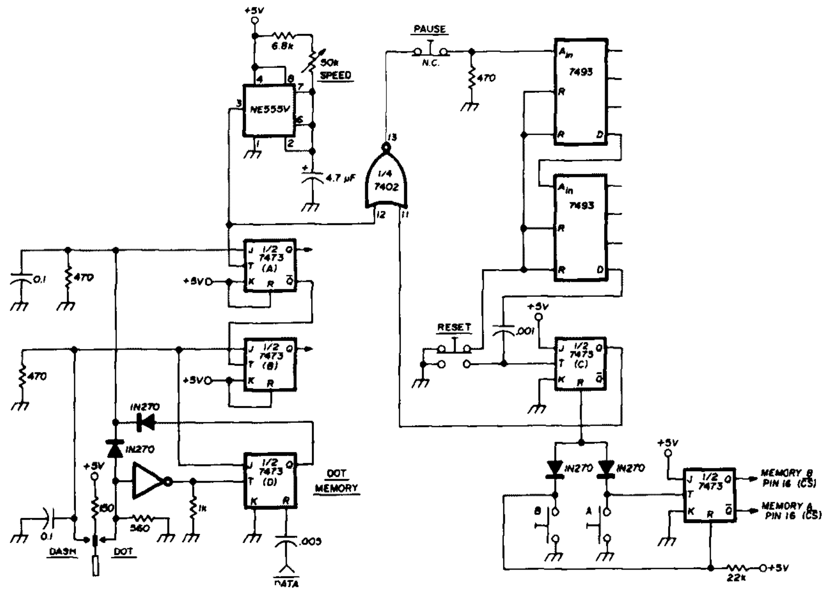



**fig. 1. Modifications to WB9FHC's memory keyer eliminate sporadic keying anomalies. Also added are a dot memory, pause and reset, and pushbutton select and start.**

A dot or dash is "remembered" and will be heard when the next clock pulse arrives. Fig. 1 shows the modified NE555 diagram.

It should be noted that this memory is not equivalent to a dot memory which prevents dots from being lost when the dot paddle is hit during a dash, as in the letter N. This feature is essential for all but the most coordinated CW operator. The dot memory shown in fig. 1 was adapted from January, 1975, *QST*[5] and makes use of the master-slave input memory. The unused half of the start/stop flip-flop, a 7473, can be connected as shown for this purpose.

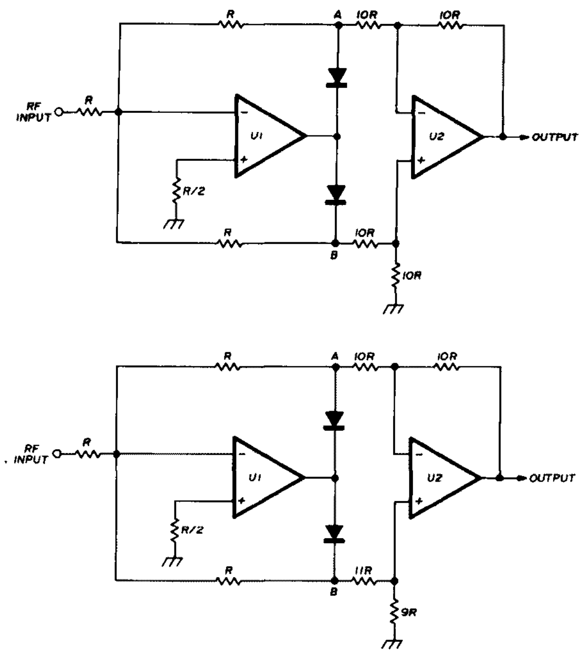
The 100 mA transformer supplied with the kit is barely adequate to supply the board's current needs, thus limiting the number of extra logic functions which can be added. The circuit changes shown reflect this problem; if a heftier 5-volt supply is used, diodes and double-throw pushbuttons can be replaced with cleaner circuitry using gates and inverters.

I have found pushbutton select-and-start to be important for contest efficiency, but needed a way to cancel a message in case of error. The circuit shown accomplishes this without additional chips. The use of an inverter could eliminate the reset-followed-by-stop action of the spdt pushbutton.

A pause feature can be added using another such switch, also shown in fig. 1. Once the timing skill is developed, this control can be used to insert repeats or additional words in a transmitted message. A more complex circuit would

1. Michael Gordon, WB9FHC, "Electronic Keyer with Random-Access Memory," *ham radio*, October, 1973, page 6.
2. Howard M. Berlin, K3NEZ, "Memory Keyer" (Comments), *ham radio*, December, 1974, page 58.
3. Howard M. Berlin, K3NEZ, "Increased Flexibility for the Memory Keyer," *ham radio*, March, 1975, page 62.
4. Howard M. Berlin, K3NEZ, "RAM Keyer Update," *ham radio*, January, 1976, page 60.

5. James H. Fox, WA9BLK, "An Integrated Keyer/TR Switch," *QST*, January, 1975, page 15.




**Schematic diagram of the improved full-wave detector with new resistor values.**

gate the memory output, eliminating the possibility of stopping the message with the transmitter keyed.

Jerry Holter, W6OCP
Menlo Park, California 94025

## diode detectors

Dear HR:

I enjoyed Hank Olson's article on "Diode Detectors" in the January, 1976, issue of *ham radio*, but would like to point out an error in his "improved" rectifier circuit of fig. 11 (shown below). The circuit as shown is not balanced, and a page or two of algebra shows that the correct ratio for the second stage lower input divider is $9R$ - $11R$ and not $10R$ - $10R$ as shown. In addition, the circuit is not new. If you let the upper divider values be $R$ and $R$ (instead of $10R$), the lower divider ratio goes to unity (i.e., it drops out) and the circuit is identical to one I published in *Electronic Design* in 1967.*

Allan Lloyd, W2ESH
Hawthorne, New Jersey

*Strictly speaking, Mr. Lloyd's criticism of the improved rectifier circuit is valid. The slight error occurs because point* A *drives a resistance of* $10R$ *(since the inverting input of U2 is a virtual ground) and point* B *drives a resistance of* $20R$ *(the input impedance of U2's non-inverting input is very high). That is, points* A *and* B *are driving different impedances by a 2:1 ratio. Since these impedances were deliberately picked to be considerably larger than the impedances driving them, little imbalance occurs (comparable with the resistor tolerances typically used in amateur projects).*

*If one wants to be absolutely accurate, Mr. Lloyd's advice should be heeded, and the circuit values changed as in the schematic above. Of course,* fig. 13 *in the original article has a small potentiometer to correct for this problem and that of resistor tolerances.*

*Until Mr. Lloyd's letter arrived I was not aware of his "Ideal Rectifier" circuit which was published previously in* Electronic Design. *To my knowledge, the circuit was the design of Dr. Nicholas Cianos, to whom I gave credit in the article.*

Henry D. Olson, W6GXN

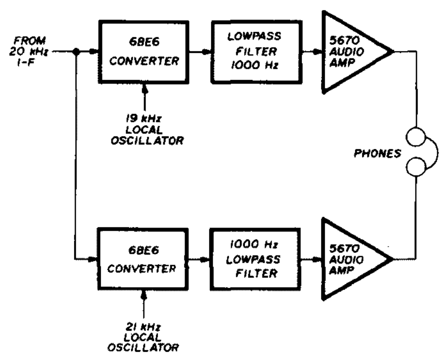
*Allan Lloyd, "Ideal Rectifier Uses Equal-Value Resistors," (Ideas for Design), *Electronic Design*, June 21, 1967, page 96. For further comments and additional features of this interesting but tricky circuit, see "letters" column of *Electronic Design*, October 11, 1967, page 46; and December 20, 1967, page 48.

## binaural cw reception

Dear HR:

The article, "Synthesizer for Binaural CW Reception" in the November, 1975, issue of *ham radio* was particularly interesting because, back in the early 1950s as W1OPN, I built an outboard unit for my homebrew receiver to achieve a similar effect to the one author Hildreth describes. In my case the center frequency from the narrow band i-f produced equal 1000 Hz notes in each ear. Tuning through the signal, from left to right for example, resulted in a low frequency beat note (left ear) rising to 1000 Hz (both ears) and then dropping in frequency again (right ear). As I recall, the left-to-right effect was not too impressive, but I'm sure Hildreth's state-of-the-art approach makes it most worthwhile. A block diagram of the system I used is shown below.

The really dramatic results, and the reason I write, occured when I tuned one of the well-isolated local oscillators to almost the same frequency as the other oscillator. The very small frequency difference (or was it phase) provided the strong sensation that the desired signal stood out in three-




dimensional space. The result was a true stereo sensation with "depth and presence" as the ads say, a sensation that made CW copy unique.

As I usually do, I tore the unit down and went on to something else. If I ever finish modifying my current solid-state receiver, I want to try this experiment again. Perhaps some other *ham radio* readers would like to try it too.

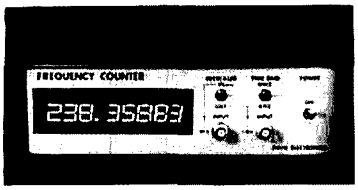
David L. Anthony, W5NOE
Columbus, Texas

# NEW products

## flexible antennas for two-meter hand-helds

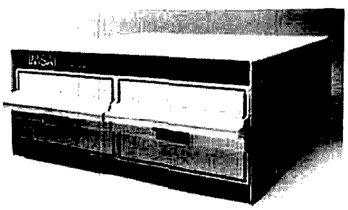
Antenna Specialists recently announced a new line of "Rubber Ducky" type antennas for two-meter hand-held transceivers. These antennas are designed for units using type-F TNC, or BNC antenna connectors, and all are protected by high quality, molded plastic sheaths. The HM-226 (TNC-type) is designed for the newer *Wilson* brand transceivers, and is priced at $11.50; the HM-227 (BNC-type) is designed for many brands, including *Genave*, and is priced at $9.50; and the HM-228 (F) is designed for the older *Wilson*, *Tempo*, and other brands, and is priced at $8.75. For additional information write the Antenna Specialists Company, 12435 Euclid Avenue, Cleveland, Ohio 44106.

## 500 MHz frequency counter




Davis Electronics recently announced its new CTR-2 Frequency Counter with prescaler that covers 500 MHz with an accuracy of 0.00002 per cent. Features of the new counter include a full 8-digit display, large 0.3 inch (7.6mm) LED readouts, automatic decimal point placement, 1-Hz resolution, high input sensitivity, and automatic input limiting. The CTR-2 has selectable gate times of 1 millisecond and 1 second, with provision for 10 seconds, and a high-stability 10 MHz TCXO time base. This counter is recommended for any frequency measuring task from 1 Hz to 500 MHz — audio through ultra-high frequencies — and is guaranteed for one year. Delivery from stock at only $349.95. For more information and *free* literature, write Davis Electronics, Dept. JJ, 636 Sheridan Avenue, Tonawanda, New York 14150.

## floppy disk



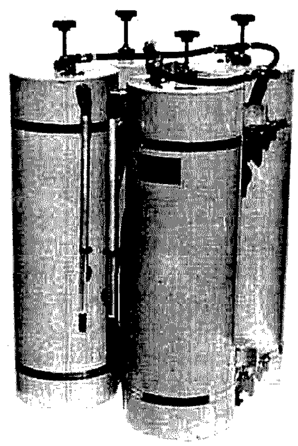
Imsai, San Leandro, California, recently announced the availability of a floppy disk drive with an intelligent interface/controller. The system is specifically designed for use with the Imsai 8080 computer, and offers the user a number of benefits heretofore not provided by other manufacturers.

The floppy disk has a capacity of 243k bytes using the IBM 3740 format. The interface/controller contains its own processor and direct access memory which operate independently but under command of the main processor of the Imsai 8080, enabling the main processor to perform other tasks while a disk operation is in process. Also, the user can change the program format of the disk by reprogramming the interface EPROM (erasable programmable read only memory) chips. Up to four floppy disk drives can be controlled by one interface/controller. Each disk can be write-protected by software control.

The disk drive comes in a cabinet with a power supply and the capacity to accommodate a *second drive.* A rack-mounted version is also available. All interconnection cables are included. The Imsai Floppy Disk Drive and Interface/Controller are $1,649 assembled and $1,449 unassembled. An additional disk drive without cabinet is $925. The Interface/Controller alone is $799 assembled and $599 unassembled. Disk Operating System software is available on diskette for $40. Also, 12k Extended BASIC with disk access capability is available.

For additional information, write IMS Associates, Incorporated, 14860 Wicks Boulevard, San Leandro, California 94577; telephone (415) 483-2093.

## bpbr circuit duplexers



Wacom Products, Inc., of Waco, Texas has announced a new line of duplexers which include the use of a new exclusive circuit developed by the company. When used with a high-*Q* filter, the *BpBr Circuit* provides superior suppression of spurious and



More Details? CHECK—OFF Page 118

sideband noise between the adjacent duplex frequencies, particularly when the duplex frequencies are close-spaced. A patent is pending on the new circuit. new circuit.

Model WP-641 consists of four 8-inch (20cm) OD cavities with the BpBr Circuit, and is designed for use with duplex stations in the 144-174-MHz band when the transmit-to-receive frequency separation is 500 kHz or more. It provides bandpass characteristics near the pass frequencies and band-reject cavity characteristics at the frequencies to be attenuated. Superior transmit-to-receive isolation is a feature of the new model.

For additional information contact Wacom Products, Inc., Box 7307, Waco, Texas 76710; or telephone (817) 776-4444.

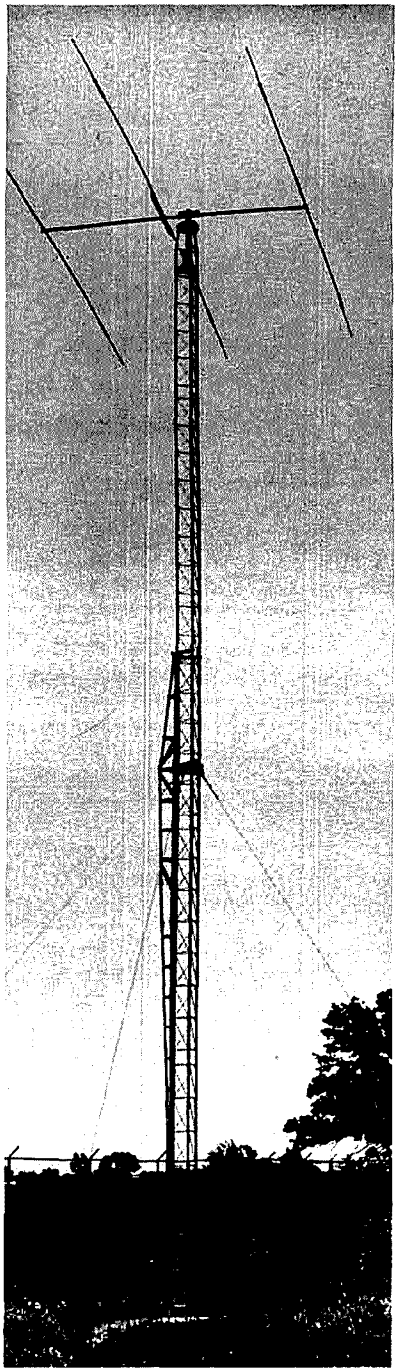
## gutter-mount mobile antennas

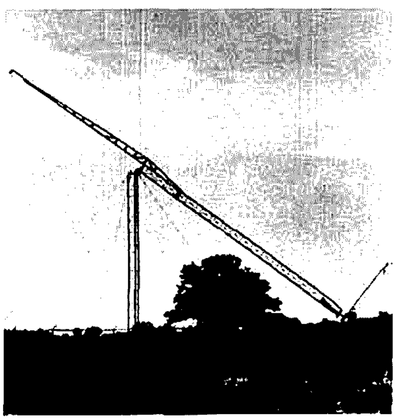
Hamline Electronics recently introduced a new line of easily-installed vhf mobile antennas featuring a stressed-aluminum gutter-mount bracket and quarter-wave elements for 146-MHz and 220-MHz applications. The bracket includes an SO-239-type connector with ten feet of RG58/U coaxial cable. A simple, plastic-tipped stabilizing screw permits the bracket to be adjusted for any vehicle rain gutter so that the antenna will be vertical. Two mounting screws threaded into a clamping plate secure the assembly in minutes to a particular automobile.

The antenna elements are made from resilient steel and are vinyl-clad for protection from weather. Each element is mounted in a non-tarnishing PL-259-type plug and sealed with epoxy to form an electrically sound, weatherproof, quickly-detachable assembly.

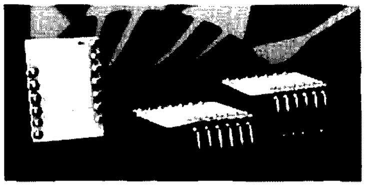
The FME-146 element for 2 meters, and the FME-220 element for 1¼ meters may be purchased separately for only $4.95 each, postpaid. The mounting bracket with one element is priced at only $15.95 postpaid. Please specify the element desired when ordering. (Pennsylvania residents please include 6% sales tax.)

For additional information, write Hamline Electronics, Box 52, Sweet Valley, Pennsylvania 18656; telephone (215) 929-8118 (preferred) to (717) 256-3017.





## active filter useful to 10 kHz



A new series of IC active filters designed for use at frequencies up to 10 kHz has been developed by National Semiconductor Corporation. Called the *AF100 series*, the new integrated circuits are intended for use in low frequency analog systems, including medical, geophysical, sonar, audio, tone signaling, modem, and feedback control systems where specific filter functions are required.

The AF100 series filters are basic building blocks that can be used to construct any filter response such as Butterworth, Bessel, Cauer, and Tschebycheff. With the addition of only four external resistors, the AF100 can be programmed for second order functions.

Lowpass, highpass, and bandpass functions are available simultaneously at separate outputs, and notch and all-pass functions are available by combining these outputs in an internal summing amplifier. If higher order systems are required, several AF100s can be cascaded. In all configurations, the *Q*, gain and center frequency adjustments are independent, requiring no iterative trimming. Other features of the new active filter series include a *Q* range of up to 500 and a frequency accuracy of either ±1% or ±2.5%. Operating power supply range is from ±5 volts to ±18 volts and supply current is 4.5 mA maximum.

The new Active Filter series is available in either a 16-pin plastic dual in-line package or a 12 pin TO-8 can for operation over the commercial temperature range of from −25°C to +85°C, and in the TO-8 can for operation over the military temperature range of from −55°C to +125°C. In quantities of 100, the commercial plastic CIP active filters with a frequency tolerance of ±2.5% (the AF100-1CJ) sells for $4.95 each.

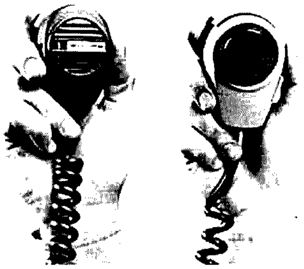
For more information write



More Details? CHECK–OFF Page 118

National Semiconductor, 2900 Semiconductor Drive, Santa Clara, California 95051; or telephone (408) 737-5000.

## shure mikes



Shure Brothers offers two new dynamic element communications microphones that set new standards for ruggedness, reliability, and transmission quality in safety, tranportation, and industrial communications applications.

The Model 524C *Ranger 2* is a transistorized unit designed to match carbon microphone-type circuits and significantly improve intelligibility and performance. Its built-in transistorized amplifier supplies sufficient output to permit the use of up to 30m (100 ft) of unshielded cable, while offering a wide frequency response.

The Model 577 *Sonobar* combines a noise-cancelling pickup pattern and a "tailored" voice-range frequency response to block out unwanted noise and boost close-up voice signals, ideal for all fixed-station and mobile applications where high ambient noise conditions are a problem.

The new Shure Model 577 Sonobar microphone is available in high or low impedance models, and a transistorized model for carbon microphone-type circuits. The new 577 series replaces the now-discontinued Shure 488 series, with the exception of the FAA-approved 488T which is still available.

User net prices of these new microphones are: Model 524C *Ranger 2*, $66; the 577A (high impedance), $76.75; the 577B (low impedance), $76.75; and the transistorized 577C replacement for carbon microphones, $84.75.

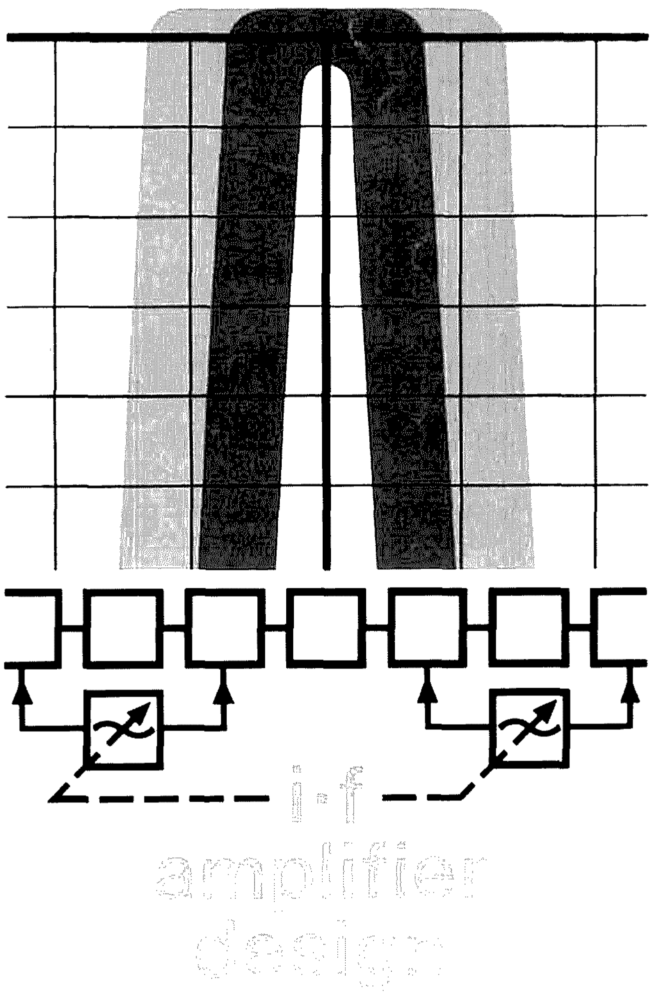
For additional information write Shure Brothers, Inc., 222 Hartrey Avenue, Evanston, Illinois 6C204.

$1.50

# ham radio magazine

## MARCH 1977





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**volume 10, number 3**



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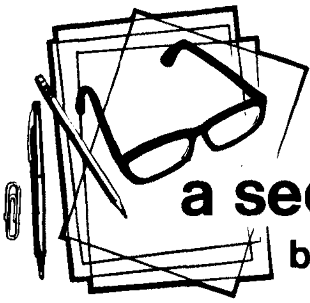


# a second look

by Jim Fisk

This year marks the 40th anniversary of the invention of the klystron in a basement laboratory at Stanford University in California – an event that marked the beginning of the practical use of microwaves. Feeble amounts of microwave power had been generated in the laboratory in the 1920s, but long-range microwave communications systems and radar required much more. By 1935 radar was already installed on naval vessels, and the Signal Corps was developing the 270 system which detected the Japanese attack force at Pearl Harbor in 1941, but the equipment was range limited and suffered from poor resolution. Microwaves seemed to be the answer, but how to produce them was a question which remained to be answered.

Russell Varian conceived the klystron – the first practical microwave generator. Dr. William Hansen, a physics professor at Stanford who was working on an oscillator for a linear accelerator, calculated the formulas for its design. In August, 1937, Sigurd Varian finished assembling the first tube, and the trio watched in amazement as the very first surge of voltage set it generating waves as short as 10 cm (3000 MHz), one-tenth the length of the shortest radio waves then used outside the laboratory. To demonstrate their success they made a dipole from two short pieces of wire and soldered it to the base of a flashlight bulb – when they held the dipole in the path of the tube's output, the bulb glowed brightly. The first klystron was delivering a watt of power!

This was one of the first cases in the history of technology development where a device based on calculations worked the first time it was turned on. The achievement was widely acclaimed in the scientific press, and broad uses were forecast for its future. The Sperry Gyroscope Company moved quickly to back the development of the klystron for high-resolution radar systems, the Bell Telephone Company saw it as the heart of a transcontinental relay network, and others considered it for coast-to-coast television transmission. The rest is history – a $100 investment in 1937 spawned a whole new industry which now has sales measured in the billions of dollars.

Before the war, Amateurs gave little thought to the frequencies above 400 MHz because there was nothing in the way of practical equipment available, but when the Amateur microwave bands were made available on November 15th, 1945, it was a different story – military surplus klystrons were inexpensive and many Amateurs had been exposed to the microwave art during the war. Rube Merchant, W2GLF, and Art Harrison, W6BMS, were the first to use the new medium when they put 2K23 klystrons and crystal mixers to work on the 5300-MHz band for communications over a 5-mile path on Long Island. In early May, 1946, W1LZV and W2JN worked over a 2-mile path in New Jersey on 10 GHz, and later that month W1NVL and W9SAD in Schnectady were the first Amateurs on 21 GHz – best DX was 800 feet! It was to be another year before 3300 MHz was opened for the first time, and then only because W6IFE built two complete stations and had W6IZM operate one of them.

In the mid 1950s members of the San Bernadino Microwave Society developed the polaplexer system and proceeded to break most of the microwave distance records. They later phase locked the klystron to a crystal reference to increase range, but little has happened in the past ten years. Some microwave enthusiasts have replaced their bulky vacuum-tube power supplies, modulators, and receivers with solid-state gear, but klystrons still provide the output power when modern solid-state sources could do the same job with a lot less hassle. Although Amateurs in England have been using Gunn diodes on 10 GHz for several years, for some reason the idea has never caught on in America. Device availability is certainly not the problem – Amateurs have shown time and time again that they can obtain hard-to-get components if they really put their minds to it (remember the parametric amplifier, 432-MHz varactor multipliers, and low-noise transistors for 2304 MHz).

I think it's time for Amateur microwavers to take a close look at their klystron systems and consider the solid-state alternatives. Gunn diodes are being manufactured by the thousands for 10.525-GHz speed radar systems so cost is low, and Microwave Associates has recently introduced a 20 mW *Gunnplexer* for transceive operation on the 10-GHz Amateur band (see page 86). In light of WARC 79 we need a good deal more activity on the Amateur microwave bands – perhaps the new *Gunnplexer* will provide the impetus amateurs need to try this part of the spectrum.

**Jim Fisk, W1HR**
**editor-in-chief**

# presstop

WARC ADVISORY COMMITTEE FOR AMATEUR RADIO met January 25 and prepared new and strengthened supporting arguments for new HF bands at 10, 18, and 24 MHz. Even though the FCC's Notice of Inquiry had not included those previously-requested bands, the group felt they were important enough that another strong plea was called for.

The Committee Endorsed, for the most part, the remaining Amateur assignments proposed in the Notice of Inquiry. The 15-meter shift down to 20.7 MHz was one notable exception, but this appears to have been worked out with the Maritime group to put the band back where it presently is.

Considerable Discussion took place concerning the microwave assignments, and a strong recommendation emerged urging restoration of 48-50 GHz (or a comparable slot) to the proposed Amateur spectrum.

THE AMATEUR RADIO MANUFACTURER'S ASSOCIATION is the name chosen for a self-governing body of radio equipment manufacturers formed at SAROC this January for the purpose of promoting Amateur Radio, promoting high ethical standards among its membership and throughout the industry, encouraging legislation and rule-making favorable to Amateur Radio, and supporting its members' activities through market research.

ARMA's Formation Came About through a Dentron-sponsored meeting of linear amplifier manufacturers who initially met to discuss the linear amplifier issue. Finding that it had greater potential then the single issue originally addressed, ARMA elected Dennis Had of Dentron as its President and began making plans for meetings with members of Congress, representatives of the FCC, and other government officials.

This Broad-Based Trade Group invites membership applications from manufacturers and importers marketing in the U.S., as well as suppliers, publishers, associations, and others interested in furthering the group's aims.

ARMA's Recent Washington Trip was rated "highly successful" by both the visitors and those they came to see. The group sat down with all three FCC Commissioners they had appointments with and the Commissioners welcomed the ARMA delegation as representatives of the legitimate Amateur industry and expressed concern for that industry, the small businesses involved in it, and — last but not least — for the Amateur fraternity.

Senator Barry Goldwater — K7UGA — spent a good deal of time with the ARMA representatives. The visitors also held highly informative meetings with Ray Spence, Bob Luff, and other key FCC personnel during their two day stay.

"PERSONAL RADIO DIVISION" is the new handle adopted by FCC's Amateur and Citizens Division. The name change is just a name change, though it could eventually reflect future concern with other general public radio such as marine and aircraft.

PERSONAL DELIVERY OF AMATEUR APPLICATIONS to Gettysburg gives the applicant no advantage and can tie up FCC personnel who could otherwise be working on the Amateur license backlog.

Applications For 1X2 Callsigns May Be Submitted before the official acceptance date to insure they'll receive prime consideration. For example, Extras licensed before July 1, 1976 could mail their applications any time after March 15 to be sure they'd reach Gettysburg before the April 1 acceptance.

"INSTANT UPGRADE" will finally become a reality and give the Amateur who successfully passes an exam at an FCC Field Office his new privileges immediately. The temporary authority is to be issued on the spot by the examining Field Office, and will be good for 90 days or until an upgraded license is issued. It will require the user to indicate his "interim" status with a special identifier, added to his call when he's using his newly won privileges, will probably go into effect in the next few months.

ALTHOUGH SETTLEMENT OF THE 900 MHZ VS 220 MHZ issue for Class E Land Mobile (CB) may still be months away, more and more indicators seem to be pointing to 900 MHz. It's also becoming apparent that the FCC may no longer look with favor on the placement of Amateur and CB Personal Radio Services adjacent to one another, wherever the frequency assignments may be located.

OSCAR 6'S TELEMETRY indicates a battery cell has failed, and there's fear other cells may be close to failure after more than three years in space. With rumors going around that LANDSAT (and OSCAR 8) won't be launched until the year's end, OSCAR users must be especially gentle with both the ailing Amateur Radio birds.

VHF/UHF ADVISORY COMMITTEE was established by the ARRL Board of Directors at their January meeting in West Hartford. The new VUAC is to work on spectrum utilization, ATV, Satellite and EME communications, work closely with the VRAC.

First Formal Assignment for the VUAC is a 1215 MHz bandplan.

# i-f amplifier design

## Further comments by a recognized authority on the optimum design of shortwave receivers

Many articles have been published during the past few years on receiver design with emphasis on the front end, but few efforts have been made to exploit the possibilities of good i-f amplifier design. Such design involves several areas: proper i-f selectivity, low distortion, a-m and ssb detectors, agc control, and noise blankers. This article presents some methods of designing these circuits to obtain a high-performance i-f system.

### noise blankers

Very few noise-blanker circuits for multipurpose applications have been published. Probably the best noise blanker ever built is the circuit used in the Collins KWM2; however, since this design is based on tubes, it's no longer considered within the state of the art. While the principle of the Collins noise blanker avoids difficulties of crystal-filter ringing, it does produce some distortion because of the circuit used for gating in the i-f path.

The principle of the Collins noise blanker is based on the idea that noise pulses originating either from ignition or other man-made noise are of fairly short duration and therefore, have enough energy, even at frequencies above 30 MHz, to produce radio-frequency interference. Collins uses a multistage 40-MHz amplifier that brings the noise pulses up to a suitable level for detection in a peak detector, and the dc voltage thus derived is used to disable the receiver rf front end and i-f, including the i-f

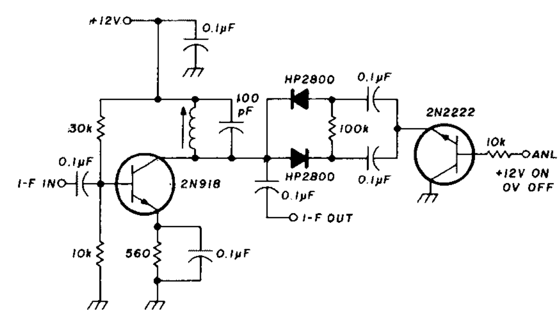



fig. 1. Rf noise limiter with self-adjusting action.

filters, during the detection of the pulses. Since some of the pulses do not have a rectangular response, the gate circuit for switching remains for a few milliseconds in a state between on and off, during which most of the noticeable intermodulation occurs.

Noise blankers in present amateur transceivers measure the noise pulses in the i-f bandwidth, and rf signals, such as CW key clicks, can't be distinguished by

By Ulrich L. Rohde, DJ2LR, 52 Hillcrest Drive, Upper Saddle River, New Jersey 07458

the circuit from other interference. Therefore all these circuits suffer from added distortion.

Fig. 1 shows a noise blanker published in the *ARRL Handbook*, which is supposedly derived from TV receivers and suppresses only certain types of noise pulses. A detailed discussion of this circuit is found in reference 1. Fig. 2 shows the principle of an i-f noise blanker that derives its information from the i-f path. This circuit, in my opinion, however, is only a compromise; however, it is found very often. Fig. 3 shows a circuit that takes advantage of the idea on which the Collins noise blanker is based and has the additional feature that it can also be used on the 160-meter band for suppressing Loran pulses. Its design is based on a suggestion by Siemens, who are the makers of the IC used in this circuit.

The TCA440 IC, to the best of my knowledge, has no equivalent American replacement. It contains a complete a-m receiver. Because of the wideband application, oscillator stability is no problem, and a fairly simple single-conversion receiver can be built to convert the arbitrarily chosen input frequency of 50 MHz to a 2-MHz i-f, where the noise pulses are then amplified to a suitable level for detection. In addition, a 2-MHz input is provided, which can be used when operating in the 160-meter band.

The IC output, which can be monitored by a special test output, is fed into a limiter and amplifier using an fet and a bipolar transistor. The TTL output is then fed into a 74LS122 IC for pulse processing, and a TTL-level noise-blanking output is available.

This circuit has substantial advantage over the Collins noise blanker but does not solve the problem of suitable

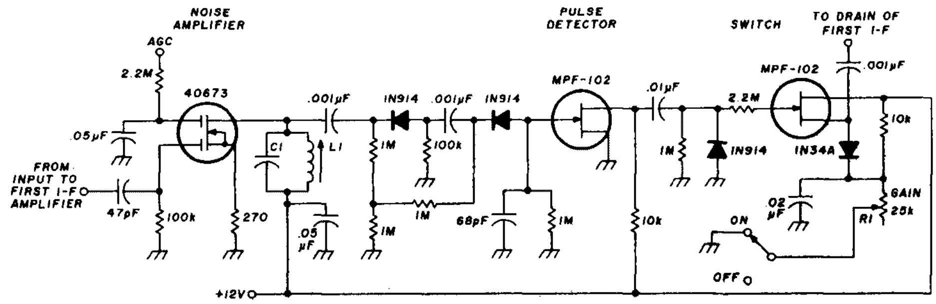



fig. 2. Schematic of a noise-blanker that derives its information from the i-f path.

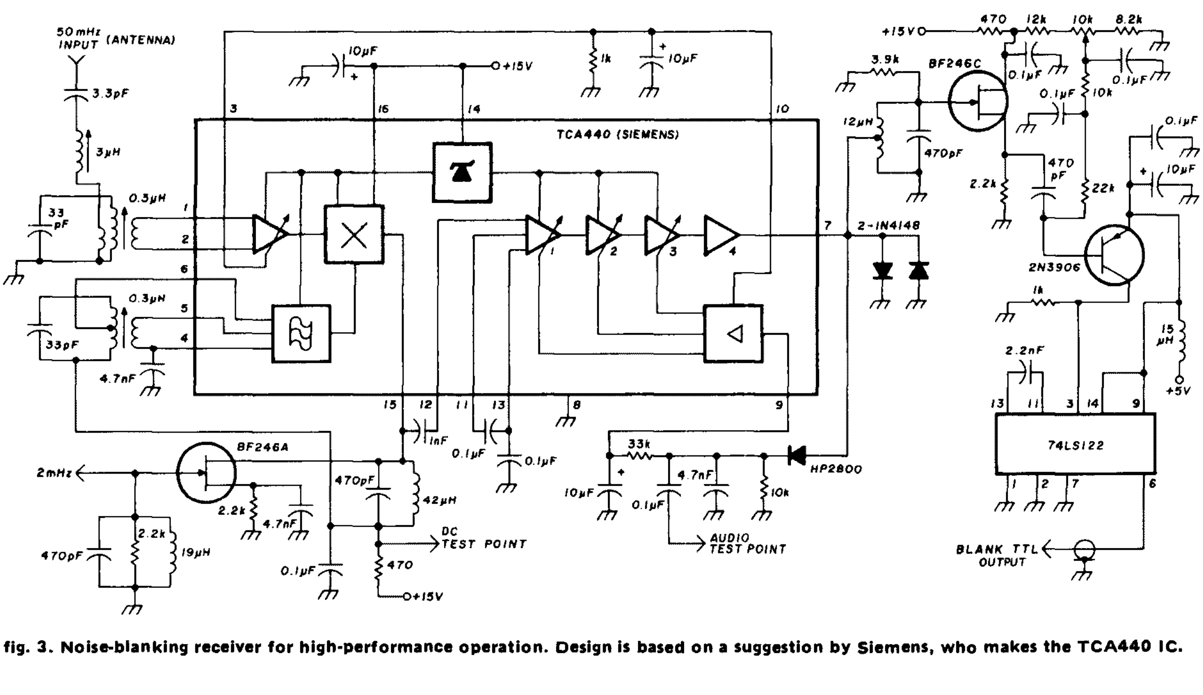



fig. 3. Noise-blanking receiver for high-performance operation. Design is based on a suggestion by Siemens, who makes the TCA440 IC.

gating. **Fig.** 4 shows an rf input stage described earlier,[2] which has a suitable push-pull arrangement that accepts these blanking pulses without producing the distortion found in most circuits. A similar gating circuit had been proposed by Swan in their SS200 series; however, since the noise-blanker circuit was extremely critical, it did not operate too successfully.

## input selectivity

For reasons explained in reference 2, selectivity should be used immediately following the first mixer to prevent overload of a possible second mixer and to avoid spurious signals. General-coverage receivers require the first i-f to be higher in frequency than the highest frequency of reception, and i-fs between 30 and 120 MHz have been used. As discussed in reference 2, the selectivity is provided by vhf crystal filters, which have 4-8 poles and are commonly called cross-modulation filters. They are available either in discrete components or as monolithic filters. Probably the most popular is the 75-MHz monolithic filter made by Piezo Technology in Florida.

Because of the losses of this filter, the stage immediately following the filter must have an extremely low noise figure and simultaneously a high intercept point. A good choice for this requirement is the circuit shown in fig. 5, which uses a neutralized field-effect transistor to obtain stability and low noise figure through rf feedback. Such a stage has about 10 dB gain. A second i-f stage with presettable gain can be used ahead of the second mixer. Because of the voltage step-down transformer, overall gain is kept very low; however a large, highly distortion-free agc range is possible.

The main selectivity is always achieved at lower frequencies. While the commonly used frequency for vhf/uhf receivers is 21.4 MHz, most attempts to make ssb or CW filters at these high frequencies have failed. For CW and ssb, crystal filters are available at frequencies of 10.7 MHz, 9 MHz, 8.8 MHz, 5.5 MHz, and 1.6 MHz, while the intermediate frequencies of 525 kHz, 455 kHz, 200 kHz, and 100 kHz take advantage of available mechanical filters. Some receivers still use LC filters at 50 or 30 kHz, but the shape factors obtainable are not very impressive.

**Variable-bandwidth** system. In military receivers that require a substantial number of bandwidths, seven or more individual filters may be required, which makes the i-f portion both bulky and expensive. To overcome this problem a double mixing scheme, as briefly discussed in reference 2, permits quasi-continuous selection of bandwidths between 75 Hz and 6 kHz at a constant shape factor of 1:1.6, independent of the bandwidth setting. **Fig.** 6 is a selectivity curve of a filter that provides *quasi-continuous selection of i-f bandwidths*.

An additional advantage of this system is that ringing from mechanical or crystal filters of very narrow bandwidths is dramatically reduced. The filter is based on a) a dual-mixing system using four mixers and two ganged local oscillators to vary bandwidths, and b) two high-grade 30-kHz lowpass filters to provide the steep slopes. **Fig. 7** shows the block diagram of this circuit.

By correct choice of oscillator frequencies and, in one case, the use of sideband inversion, the two sidebands of the incoming i-f signal can be shifted toward or away from the sharp cutoff characteristics of two identical 30-kHz lowpass filters. After the two sharp edges have been imposed on the i-f signal, it is then converted back to the original i-f center frequency, which is 300 kHz. In effect, a variable bandpass filter is synchronized with two fixed lowpass filters. The basic relationship between the oscillator frequency, intermediate frequency, lowpass frequencies, and the bandwidths are:

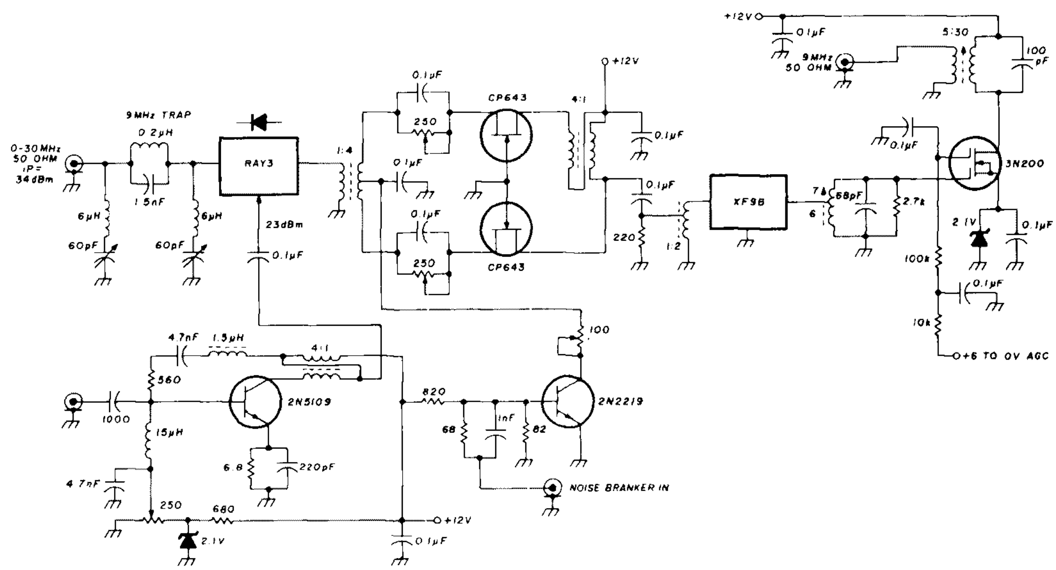



**fig. 4. Rf-input stage for a high-performance receiver. Push-pull arrangement accepts blanking pulses without producing the distortion found in most circuits.**

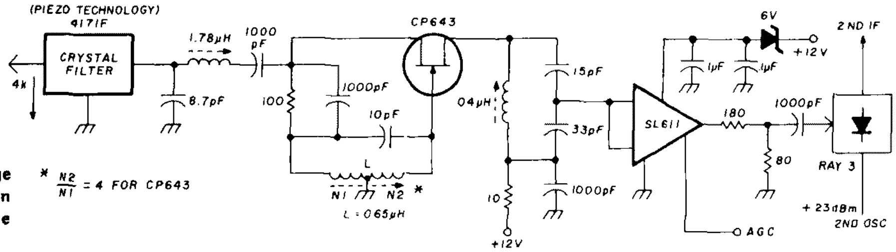



**fig. 5. Low-noise neutralized input stage with crystal-filter matching and an output stage for i-fs in the vhf range. The filter is by Piezo Technology Co.**

$$f_1 = f_{i\text{-}f} - f_g + f_b$$
$$f_2 = f_{i\text{-}f} + f_g - f_b$$

where

$f_{i\text{-}f}$ = the intermediate frequency
$f_g$ = the cutoff frequency of the lowpass filter
$\pm f_b$ = the required filter bandwidth
$f_1$ = the first oscillator frequency
$f_2$ = the second oscillator frequency

As the oscillator frequencies change (in opposite directions), the filter bandwidth narrows. By independent adjustment of the oscillator frequencies, various asymmetrical filter characteristics can be set up.

The advantages of this quite complex filter derive from the fact that it is possible to build high-grade lowpass filters at 30 kHz using crystals for improving the selectivity; therefore, they have an almost ideal slope characteristic. This slope is almost exactly the same for all bandwidths. The advantage of being able to select the optimum bandwidth for all signal conditions does not have to be stressed.

**Bandpass tuning.** A technique that has become quite popular is "bandpass tuning." This method of shifting the i-f bandwidth toward the carrier center frequency was first used by Collins. It should not, however, be confused with the technique described above to achieve a variable bandwidth, since the absolute bandwidth remains the same.

The bandpass tuning *cannot* improve the selectivity, although it sometimes appears as if this happens because the audio pitch has changed and sometimes the filter center frequency is moved into the opposite sideband. This popular technique is used mainly in receivers where the designer wants to avoid the expense of crystal filters with better skirt selectivity. Bandpass tuning as presently used does not only leave the selectivity the same as it was before, but because of the additional oscillators required, receiver performance is degraded. The additional mixing stages make the receiver more vulnerable to overload, and the limited amount of shielding does not suppress oscillator harmonics sufficiently, so that an

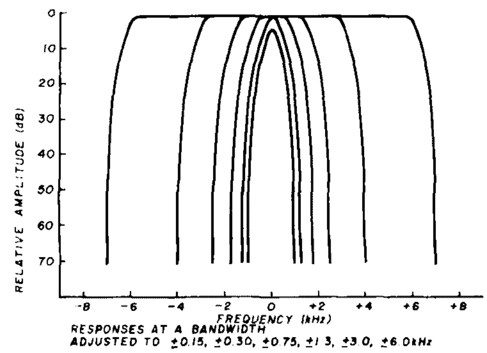



**fig. 6. Frequency response of a quasi-continuous-bandwidth i-f system.**

enormous amount of birdies are produced. An example for these tradeoffs is the otherwise very good R4C receiver made by Drake.

The use of the R4C on 10 meters, together with converters for 2 meters and higher frequencies, is very limited because of the large number of in-band birdies. **Fig. 8** shows a typical scheme to obtain bandpass tuning. This scheme is designed so it can work with a transceiver, as mentioned in my article[3] on various synthe-

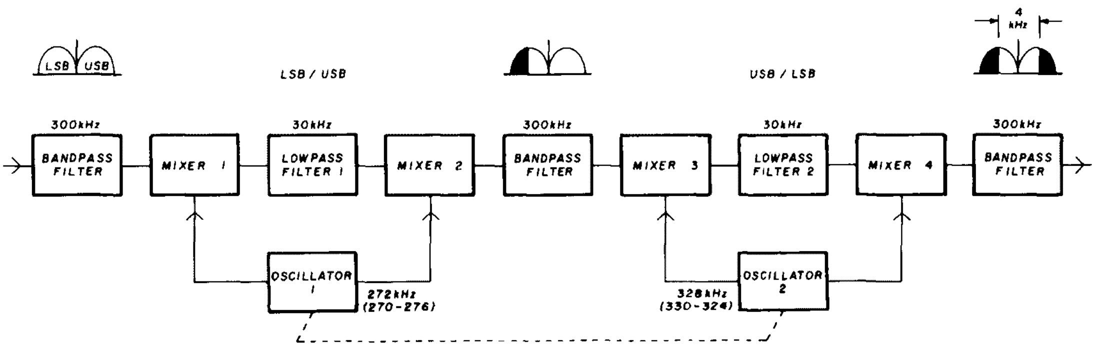



**fig. 7. Block diagram of a variable bandwidth i-f system.**

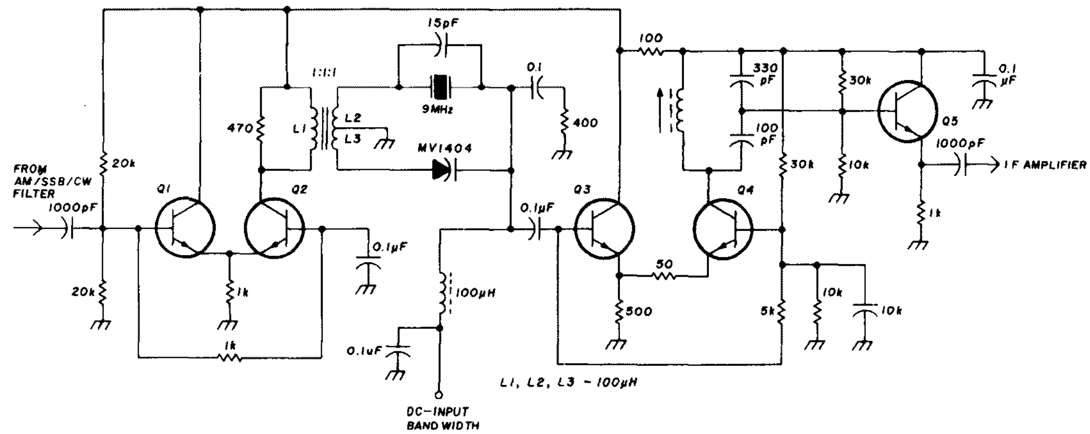



**fig. 9. Crystal notch filter for frequencies between 5-20 MHz. A half-lattice network with phasing capacitor forms the basis for this circuit.**

sizer techniques. Because of the frequency differences used in this case, the amount of spurious response is small.

While the Drake receiver has comparatively little shielding, the Signal-1 CX7, with a much more complex local oscillator, avoids some of the spurious response by selecting more suitable auxiliary frequencies. However, the two units should not be compared because the CX7 uses a 42-MHz i-f, which creates other problems.

For CW and ssb reception, notch filters can be a tremendous help. While earlier receivers used notch filters at very low frequencies; *e.g.*, 30 kHz, hardly any receivers at the new standard i-fs between 5 and 20 MHz are equipped with such a device. The solution to this problem is the design of a crystal notch filter.

Fig. 9 shows the circuit of a notch filter that can be added to practically all receivers. In designing the filter it is important to make the 3 dB bandwidth slightly wider than the bandwidth of the preceding ssb filter, since only the nulling effect (parallel-resonant mode of the crystal filter) is used.

The 3-dB bandwidth is determined primarily by the 470-ohm resistor in parallel with L1. If the resistor is made smaller the bandwidth decreases, and if the resistor is made larger, the bandwidth increases but the notch depth becomes less. For proper termination and loss compensation, the circuit is built around the CA3086 IC, using two differential amplifiers. The series-resonant frequency of the crystal must be at the center frequency of the i-f. The circuit will work well at any i-f between 5 - 20 MHz. It should, however, be pointed out that the half-lattice filter has a very poor slope characteristic and must be used in conjunction with a normal filter for ssb. A separate filter for CW should be used also. The notch depth is about 1 Hz wide. This filter arrangement, which costs less than bandpass tuning, provides added selectivity, which bandpass tuning does not, and is of much greater value.

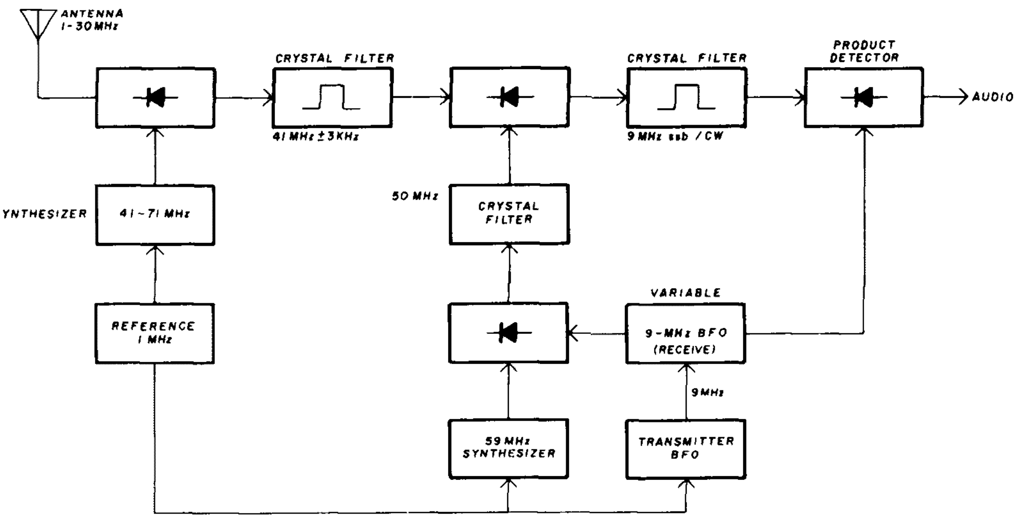



**fig. 8. Bandpass-tuning system for a transceiver.**

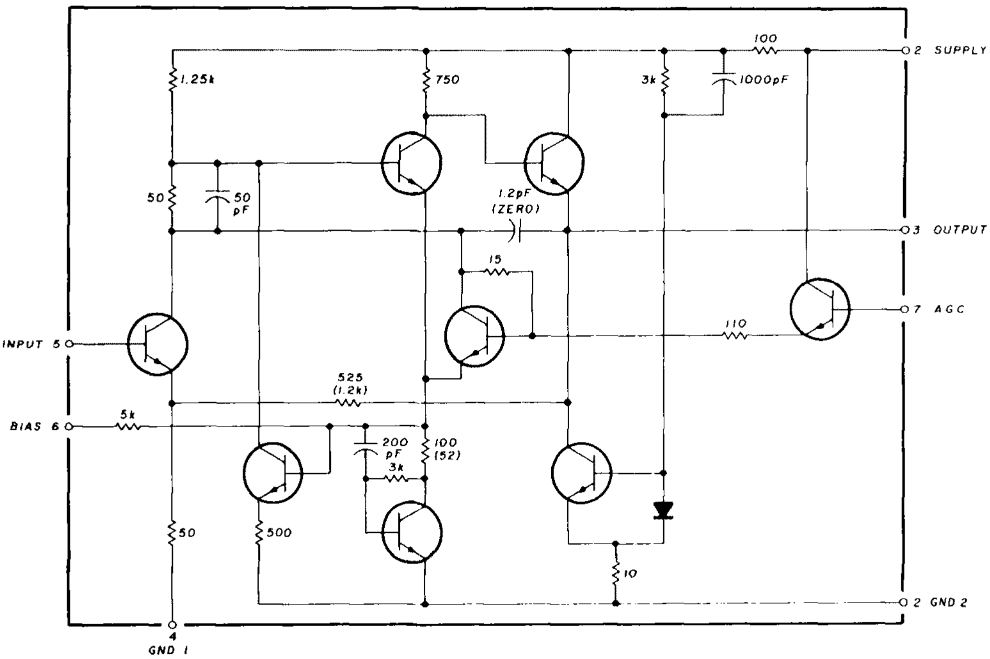



fig. 10. Schematic of the Plessey SL610/11/12 IC. Device is useful for forward agc but requires a high load impedance to avoid distortion.

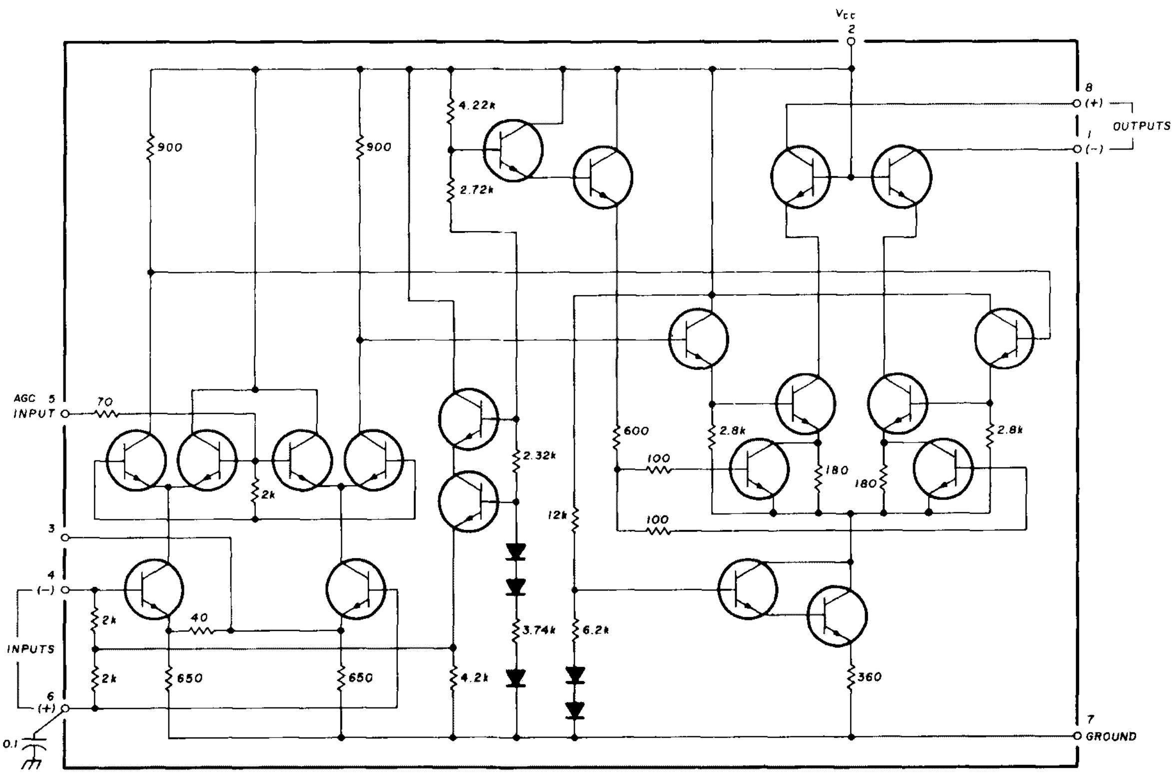



fig. 11. Schematic of the Motorola MC1349 IC. Recommended for optimum i-f circuit design using forward agc.

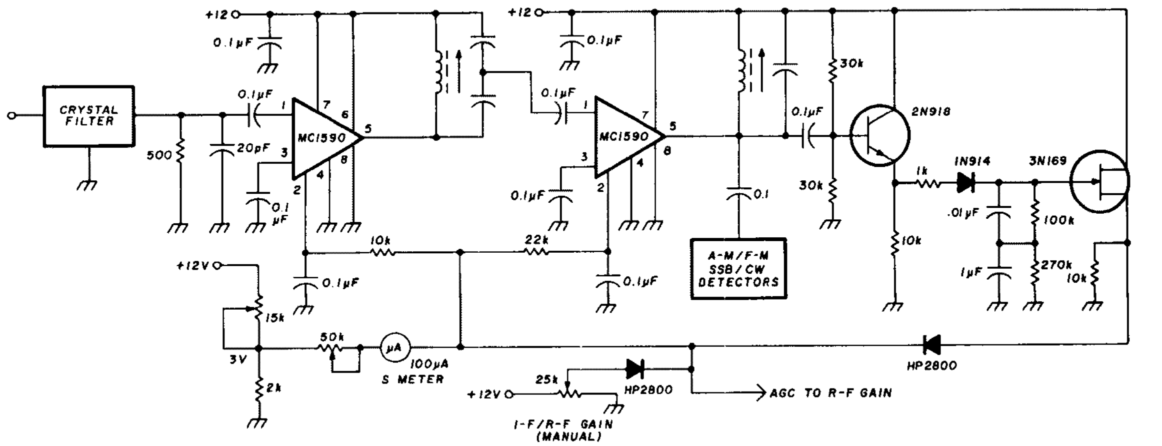



fig. 12. I-f amplifier strip with tuned circuits using Motorola MC1590 ICs. Circuits can be used for any frequency between 0.1-30 MHz.

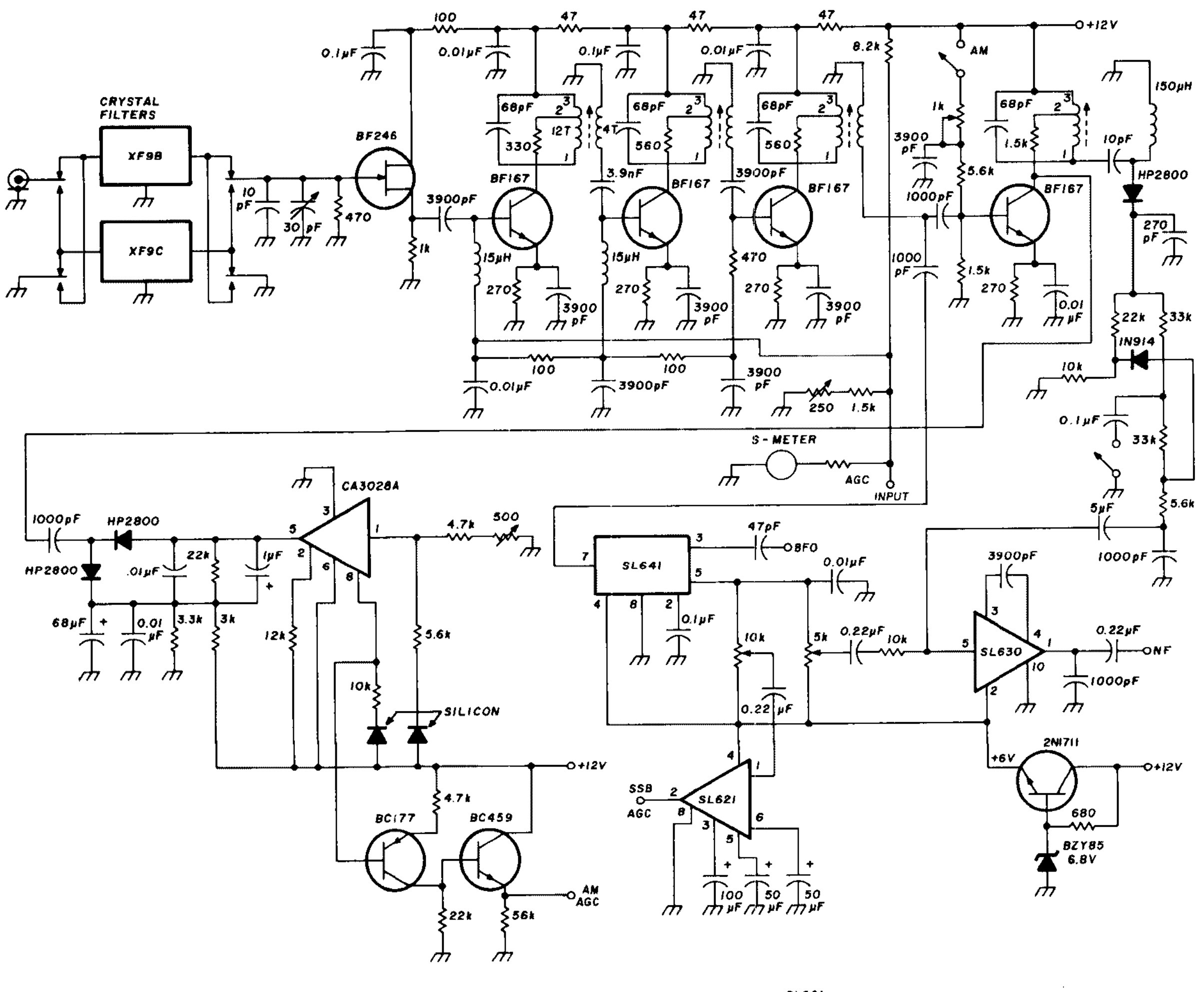
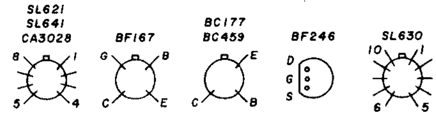



fig. 13. Complete i-f strip for a-m, ssb, or CW using discrete transistors. Forward agc is used for an ultralinear S-meter. The agc input is connected to either the ssb/CW agc output or the a-m agc output by the mode switch.



Depending on the application, the i-f stage following the main selectivity circuits can be built in various ways.

**General-coverage receivers.** General-coverage receivers differ from amateur-band-only receivers mainly by the requirement for high-performance a-m and fm detectors. Amateur-band-only receivers are tailored more for ssb and CW reception. The otherwise very interesting Plessey SL600-series IC, for example, is not suited for high-performance a-m reception because the wideband noise generated by two or more cascaded SL612s would demand very high selectivity; *i.e.*, an expensive crystal filter. Otherwise, no suitable signal-to-noise ratio is obtainable.

**Amateur receivers.** Amateur receivers in many cases use audio-derived agc to save costs. Audio-derived agc, which is not found in commercial or military receivers, has two significant drawbacks:

**1.** Inband intermodulation distortion is produced, which cannot be tolerated for many receiver uses.

**2.** When an a-m station carrier is not modulated, the agc will pump and create heavy distortion.

A fast attack time is somewhat difficult to obtain because of the limited bandwidth of the agc system. Before going into further details, it is important to take a look at the transistors and integrated circuits available for i-f stages.

In the past, the most common mistake in designing receivers was to use very little i-f gain for obvious reasons. In moderately priced receivers, which do not use extensive shielding, it is not easy to have i-f gain. Transistor circuits in this respect offer a distinct advantage over vacuum-tube circuits since the former can be designed with low impedances. Also, because the feedback capacitors, which are mostly determined by the circuit layout, are more or less the same, transistor stages permit higher gain. Some receiver designers still use discrete transistors, because they already have them, which tends to complicate the design.

**Forward agc.** Very few transistors are suitable for forward agc characteristics. A better way of doing this is to use differential amplifiers for agc. The RCA CA3028 is a suitable choice.

A somewhat more convenient way is to use ICs

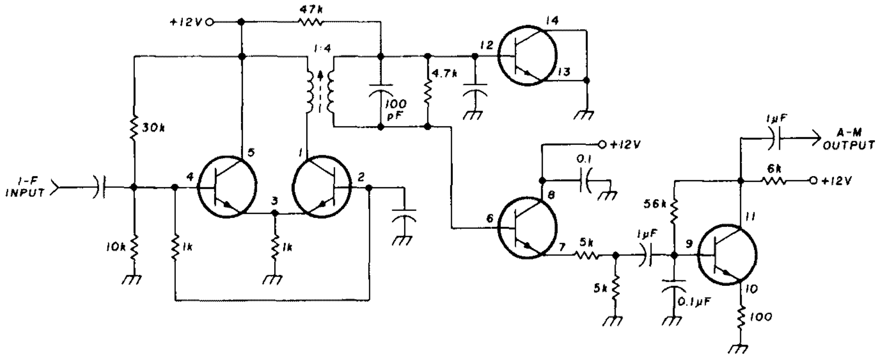



**fig. 14. High-performance a-m detector for higher frequencies. Less than 1% distortion is claimed at 99% modulation.**

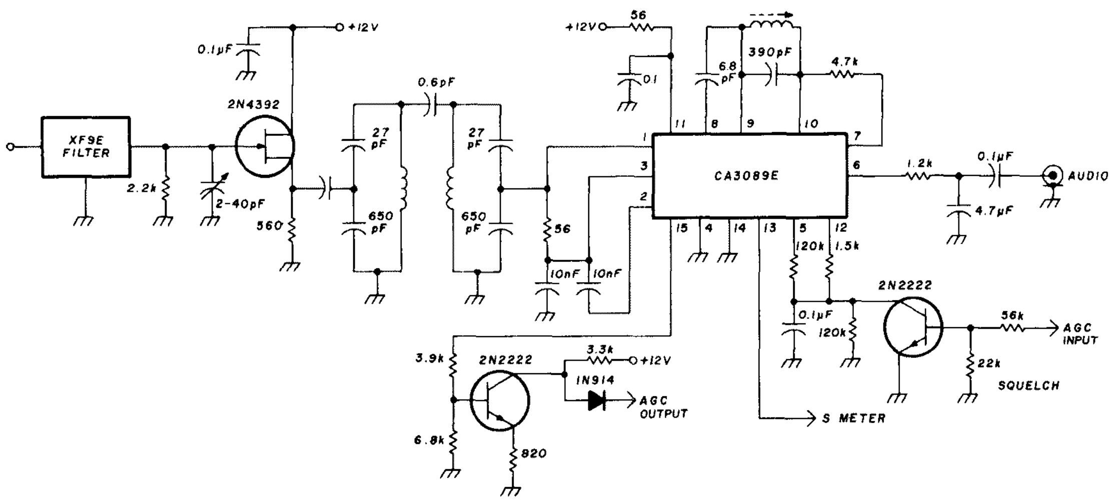



**fig. 15. Narrowband fm detector including agc and squelch circuits.**

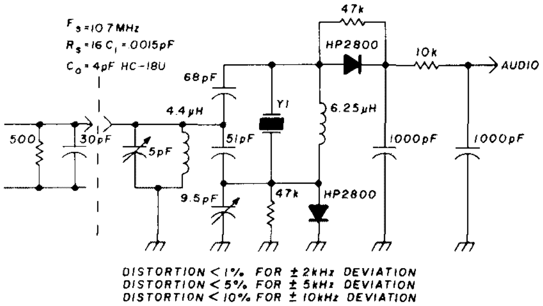



**fig. 16. High-performance cyrstal discriminator. Bandwidth is ±30 kHz with less than 10% distortion for ±10 kHz deviation. Slew rate is −75 mV per kHz. Crystal Y1 is series resonant at 10.7 MHz.**

specifically designed for this purpose, such as the Plesssy SL600 series or the Motorola MC1349/50. To understand the inherent advantages of these devices, let's take a look at their schematics. **Fig. 10** shows the circuit of the SL610/11/12, and **fig. 11** shows the MC1349 IC. The basic differences between them are:

1. The SL610 is a single-ended arrangement, which requires few external components, while the Motorola MC1349/50 series use a push-pull arrangement, which has somewhat wider applications and better dynamic range.

2. Because of the open collector, the MC1349 can be used very conveniently in narrowband applications, *i.e.*, where a-m reception is required, and still can be used with good gain at medium output impedances.

3. The SL600 series requires fairly high load impedances to avoid instability. Cascading more than two of them can become quite exciting!

National Semiconductor offers the 300 series IC for linear operation, but these amplifiers include too many functions and don't leave the designer much choice for optimum circuit design. In my opinion, the MC1349/50-series IC is the best choice, unless the full military temperature range is required or steady levels are available. In this case, the SL600 series ICs are favored.

**Fig. 12** shows a typical multipurpose i-f amplifier operating around 9 MHz, using two of the Motorola ICs with suitable agc distribution and output selectivity. If only ssb operation is desired, the tuned circuit can be replaced by 1k resistors.

**Fig. 13** shows a somewhat less expensive i-f amplifier with low-cost discrete transistors as used in some European TV receivers. Forward agc is used.

## demodulators

General-coverage receivers require a-m, ssb/CW and also fm detectors. The fm detector is required because 29.7 - 30 MHz is a popular NATO frequency range where portable fm units are being used.

**A-m detectors.** Although the a-m detector may not be all that important for most amateur purposes, I believe everyone should have a general-coverage receiver for a-m reception of foreign news services, which can be most interesting. If this is done with the usual ssb detector, the long-term stability requirements of the local oscillator and bfo are fairly high, because the carrier will otherwise introduce distortion.

High-quality a-m detection is very expensive because the rf level must be several times the diode threshold voltage to avoid distortion. On the other hand, this means that the i-f gain must be so high that a certain minimum amount of shielding, even with low-impedance, wideband circuits, becomes vital. At lower frequencies (either the 20 - 30 kHz range or the 100 - 500 kHz range) this is quite easily accomplished, because the amount of feedback is negligible. At higher frequencies, because of instability problems, special circuits should be used, which on the one hand accept very little rf drive and on the other exhibit extremely low distortion.

Probably the best circuit for this requirement is shown in **fig. 14**. This detector is basically an emitter follower with extremely little forward bias, which barely compensates the threshold voltage of the transistor. Because of temperature drift, a temperature-compensating diode is required. Although it has no gain, this circuit exhibits less than 1 per cent distortion even up to 99 per cent modulation, and is commonly called a class-D detector.

**Ssb and CW detectors**, often called product detectors, are mixers in which the suppressed carrier is reinserted. This carrier is very close, if not equal to, the i-f center frequency, which can create a problem. If the rf feedback in the i-f system is too high, radiation into the first i-f stage can cause instability and inband distortion. The best way of determining whether this problem occurs is to check the product-detector stage dc levels to ascertain whether any dc voltage offsets occur. Many good circuits

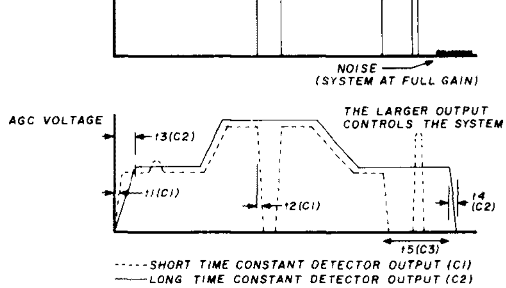



**fig. 17. Agc characteristic and pulse performance of the Plessey SL621 agc generator.**

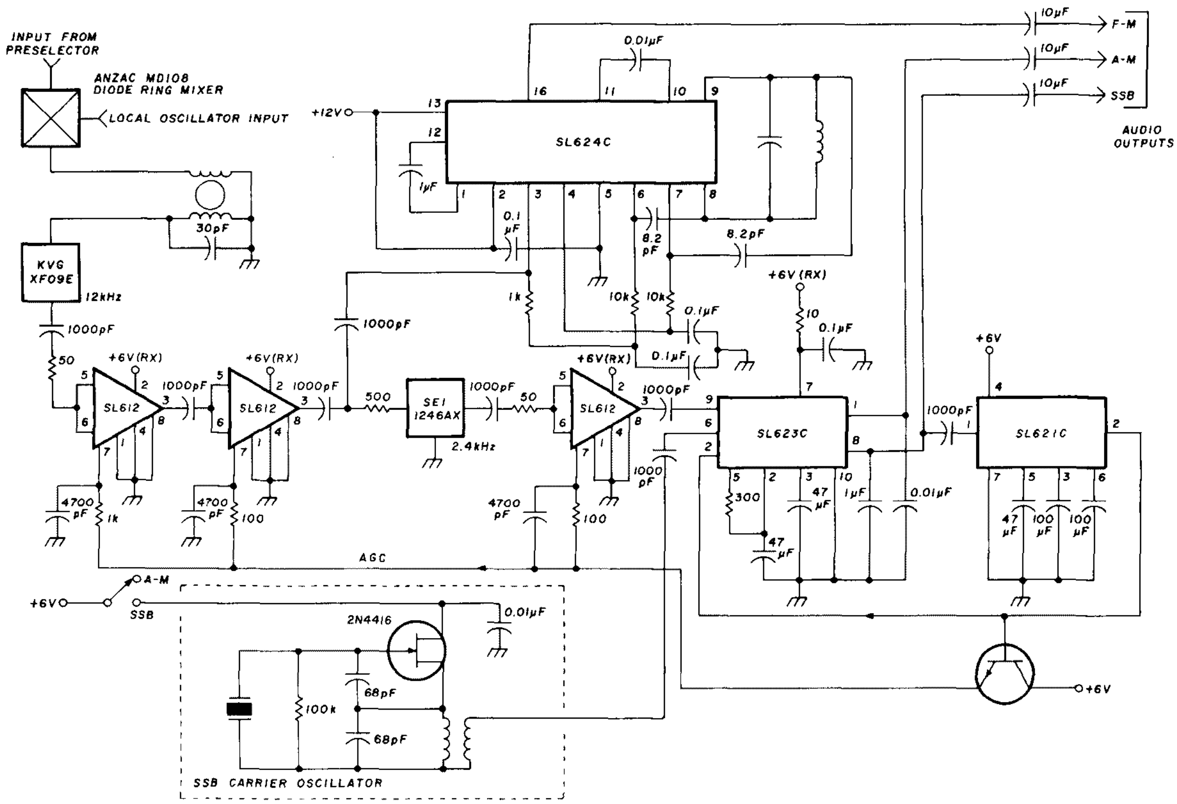



**fig. 18. I-f arrangement using the Plessey SL600-series ICs.**

are available for use as product detectors; often fancy names are used for these mixers.

The product detector is essentially a high-level mixer in which rf signals of 5 mV and higher are converted into audio signals. Lower values of rf are not recommended, because the signal-to-noise ratio then becomes problematic. Despite comments made by various authors, it is not necessary to use a four-diode ring modulator, which requires fairly high bfo signals. A popular solution is to use dual-gate mosfets or the Motorola MC1496 integrated-circuit mixer.

The dual-gate mosfet, because it is not a balanced or double-balanced switch, has no noise sideband cancellation and still requires fairly high bfo signals, while the MC1496, to operate successfully, requires an unjustified amount of external circuitry.

In my opinion, the only good choice is either the Plessey SL640 or SL641, which require few external components. Fifty mV of local-oscillator drive is sufficient for low-distortion detection, and the intermodulation distortion figures are extremely low. Detailed information on the Plessey ICs and their application are found in reference 4.

**Fm detectors.** Very few good fm detectors with little compromise are available. However, the RCA CA3089E is a unique device. It is ideally suited for amateur purposes because it offers not only excellent fm detection, high audio output, and squelch operation, but also offers a field-strength indication. The CA3089E can be used for narrowband as well as wideband fm with very low distortion, and it has been used very often. **Fig. 15** shows a recommended circuit using this IC for narrowband fm detection.

**Crystal discriminators.** In general it is not so easy to design low-distortion crystal discriminators, and most of the application data is not complete. **Fig. 16** shows a proven circuit that has ±30 kHz bandwidth. The output is ~75 mV/kHz and distortion is less than 10 per cent for ±10 kHz deviation. The crystal parameters are significant in this particular circuit, which should be terminated with a load greater than 100k.

Depending on the i-f, the crystal series-resonant frequency may vary; however, the following dynamic values should be obtained:

Series resistor, $R_s = 16$ *ohms*

Series capacitance, $C_s = 0.015\ pF$

Crystal capacitance, $C_o = 4\ pF$

The drive impedance should be 500 ohms with 30 pF in parallel.

## agc stages

Very often amateur receivers use audio-derived agc. I am not very much in favor of this method because of heavy agc pumping on a-m signals. However, it is a

convenient way of deriving agc. The overall i-f gain doesn't have to be very high, since about 14 mV of rf energy will be converted into audio voltage by the product detector, which can be used to derive the agc voltage. Regardless of how the agc voltage is derived, the time constants must be considered carefully.

For ssb and CW reception a fairly fast attack time (of the order of 2 ms) is desirable. A faster attack time has the disadvantage that the peak value of *all* noise pulses is rectified, thus blocking the receiver. A long attack time; *i.e.*, 20 ms and higher, results in unpleasant noise bursts, and some time is required before agc overshoot settles.

**Hang** agc. In earlier receiver designs agc decay time depended on the discharge of a capacitor, so that a long recovery time was required for the receiver to regain its sensitivity and selectivity. Modern circuits now use a system known as hang agc. In this system the agc values change for a given amount of time to compensate for noise bursts; thus short pauses in the signal are smoothed out. Once transmission of the signal ceases, the receiver agc voltage decreases very rapidly; therefore the receiver regains its full sensitivity and selectivity within a very short time.

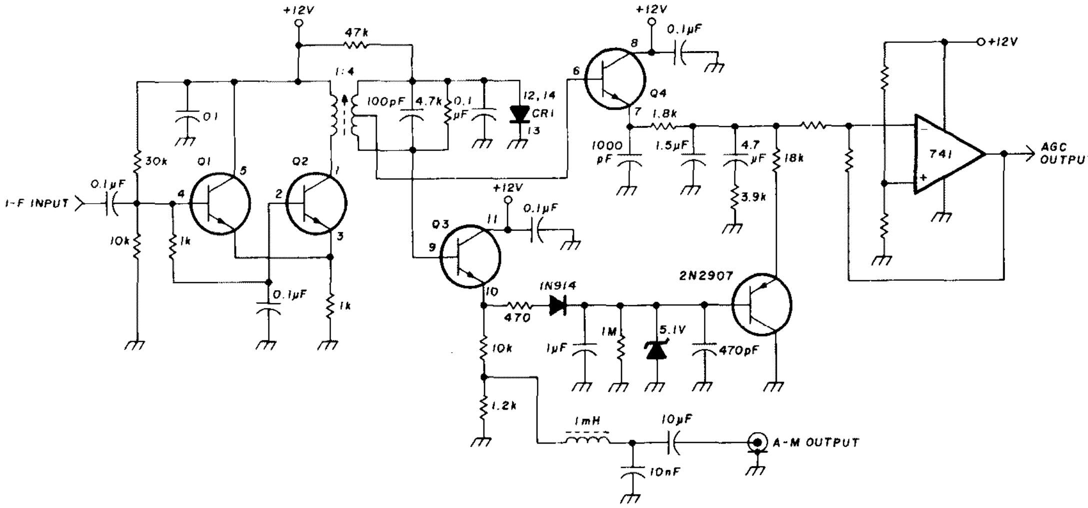



**fig. 20. Rf-derived agc system for a high-performance, general-coverage receiver. Transistors Q1, Q2, Q3, Q4 are in the RCA CA3086 IC.**

A novel approach to the hang agc circuit was made by Plessey in their SL621 agc generator, although this circuit uses audio-derived agc. **Fig. 17** shows the response of this detector; the schematic is shown in **fig. 18.** The circuit consists of an audio amplifier and two peak detectors with two different time constants which, within certain limits, permit you to select the desired attack, hold, and decay time.

Unfortunately, in this particular circuit it is not

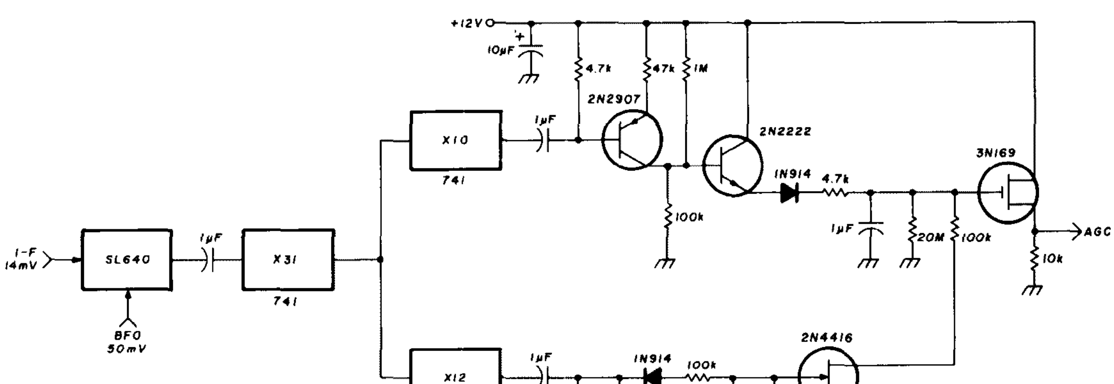



**fig. 19. Audio-derived, dual-time-constant agc system/i-f amplifier in which attack, hold, and decay time can be set independently.**

possible to reduce the attack time below 20 ms. Therefore a similar circuit was developed where all three time constants could be chosen independently. Credit for this circuit should be given to Wes Hayward,[5] although he uses split operation whereby part of the agc voltage is derived from the i-f signal and part from the af signal. The circuit shown in fig. 19 is fully audio-derived and is suggested for those who are interested only in a simple ssb/CW receiver.

The emitter-follower is responsible for the fast attack time. The high impedance input of the fet, together with the 1-$\mu$F capacitor and 10-megohm time constant, gives a 10-second maximum hold time. However, the actual hold time is determined by the auxiliary pnp circuit, which is activated after the signal disappears and which determines the actual hold and decay time.

## agc for a-m

For a-m detection, a dc voltage is available from the class-D detector described earlier. With the aid of operational amplifiers it can be brought to the required level and polarity. A suitable general-purpose agc detector for a-m and ssb/CW is shown in fig. 20 using the same principle.

Fm receivers do not need agc, although to avoid severe saturation of the i-f amplifier, which results in nonlinear phase shifts that unfortunately then become an fm-detectable signal, a certain amount of agc can be derived from the fm detector described earlier.

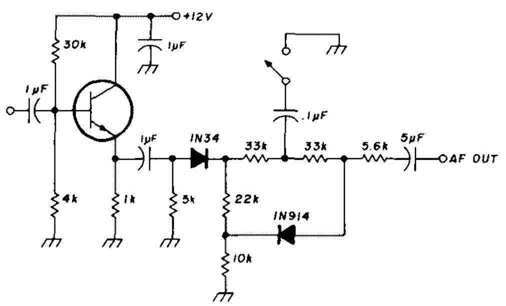



**fig. 21. Audio noise limiter with low distortion.**

## audio stages

Although not considered as part of the receiver i-f system, the audio stages deserve some mention when considering overall receiver design. Described below are two methods of enhancing receiver performance in terms of noise limiting and selectivity in the audio section.

**Noise limiter.** Noise pulses measured immediately after the first mixer can be as high as 100 times the level of the desired signal. The crystal filter reduces the amplitude of these pulses but introduces a delay, which causes ringing, or overshoot. Audio noise limiters, in my opinion, are generally worthless. However, there is one circuit that is fairly efficient. Fig. 21 shows the circuit, which I have not yet seen in the American literature. Its basic principle is similar to that of the delay line. The voltage developed across the voltage divider at the germanium-diode output is instantaneous, while the dc

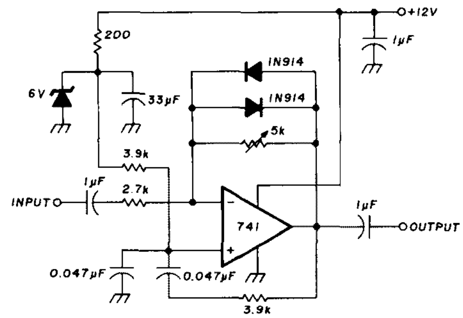



**fig. 22. Bandpass-type active filter with 800-Hz center frequency. Circuit features adjustable bandwidth and noise-limiting capability.**

voltage at the output of the circuit is delayed. If no pulses are present, and the 0.1-$\mu$F capacitor is not at ground potential, the silicon diode will have a floating voltage potential and will *not* be activated. Heavy positive pulses will charge the 0.1 $\mu$F capacitor and the silicon diode will short circuit the audio voltage. Negative pulses will disable the germanium diode directly. Therefore, a noise blanker at audio level in both directions is obtained.

**Audio selectivity.** Additional selectivity can be obtained with suitable audio stages. A common way of doing this is to use active filters. Either lowpass or bandpass arrangements can be used. It is very important to use a design in which the three parameters can be chosen independently: gain, bandwidth, and operating frequency. When not properly designed, these circuits act like a Q multiplier, which has a sharp, needle-like frequency response and little skirt selectivity.

One way out is to cascade some lowpass filters; however, the signal-to-noise ratio can be degraded. Fig. 21 shows an active bandpass filter I've used in the past. It is very well suited for CW reception, and the two back-to-back diodes act as a very convenient noise limiter.

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**ham radio**

# improved digital afsk

An answer to the need for high-quality peripheral equipment in amateur RTTY and remote computer applications

For many years the biggest problem for RTTY enthusiasts has been in mating teleprinter units to amateur equipment. As far as the receiving converter or TU is concerned, much effort has been expended to maximize performance, since the result of such effort is obvious. The transmitting converter, or afsk, however, has been generally neglected. Few really new designs have been developed since the inception of RTTY in the 1940s, and practically all have suffered from shift clicks, shift bounce, slow shift, instability, loop noise, and voltage sensitivity. Anyone who has built and attempted to use a phase-shift type afsk[1] is well aware of these problems. Some attempts have been made to incorporate digital innovations,[2,3,4,5] but these still had some drawbacks. Invariably, parts count and current drain were high, multiple power supplies were needed, portions of the design were redundant, and others neglected or omitted. The design presented here is an attempt to correct these shortcomings.

This article offers improved quality in RTTY hardware for amateur use. Also, with the expansion in data communication and computer experimentation, much need exists for high-quality interface equipment. The circuit described here is also designed to accommodate tone frequencies used in remote computer applications.

## objectives

The design criteria to be met by this circuit are:

1. Minimize the number of special or critical components.

2. Maximize stability and eliminate the need for sophisticated equipment for setup and test.

3. Minimize distortion and switching transients.

4. Incorporate a keyboard debounce circuit to reduce mechanically induced noise.

5. Produce sine-wave output at moderately low impedance and at reasonably high level.

6. Operate from a single unregulated power supply at low current.

7. Obtain complete electrical isolation of afsk from the loop to minimize ground-path noise coupling.

8. Minimize adjustments for balance and level.

9. Make loop connections polarity insensitive.

To see how these objectives have been met, and to understand the operation of the circuit, refer to **fig. 1** while following the technical description.

## technical description

U1C and U1D are two independent continuous-duty, Pierce-type, crystal oscillators running at 2048 (or 4096 for 1070-1270-Hz data shift) times the desired mark and space frequencies respectively. R1 and R2 set the dc bias on the gates. The unused inputs are tied to $V_{cc}/V_{dd}$ through R3 and R4 to ensure logic 1 on the inputs. C1 and C2 ensure reliable oscillator start up. The outputs of these oscillators are buffered by U1A and U1B and are fed to the selector gate, U2. Both oscillators run continuously to eliminate startup delay. The U2A output is either the mark or space harmonic as determined by the state of U3. If U3, pin 3, is high (loop closed, loop current flowing) then U2C is enabled, gating the mark tone harmonic. If the loop opens, U3, pin 3, will be low, enabling the space harmonic through U2D. (The action of U3 will be described later). A maximum of one cycle of distortion is introduced by this switching action, but this distortion appears at 2048 (or 4096 for 1070-1270-Hz data shift) times the mark or space frequency, and thus only contributes about 0.05 per cent maximum distortion to the output tone shift.

U2A output is fed through an eleven- or twelve-stage ripple counter, U4, whose output is a symmetrical square wave at the desired mark or space frequency. A CD4020 14-stage divider is used, but the Q10 or Q11 output is used to allow oscillator operation near the frequency of maximum stability. The divider output is fed through a single second-order active bandpass filter, U5, whose Q is set at about ten, gain at about 0.2, and center frequency between the desired mark and space

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**table 1. Changes required to the schematic for operating frequencies other than standard mark and space.**

| jumper | afsk use | Y1 (MHz) | Y2 (MHz) | R5 (ohms) | R6 (ohms) | R7 (ohms) | mark (Hz) | space (Hz) |
|---|---|---|---|---|---|---|---|---|
| 15-A | RTTY | 4.352 | 4.700 | 270k | 120k | 1k | 2125 | 2295 |
| 15-A | ASCII comp | 4.557 | 4.147 | 270k | 120k | 1k | 2225 | 2025 |
| 1-A | ASCII term | 5.202 | 4.383 | 6.8k | 27k | 250 | 1270 | 1070 |

all resistors 1/4 watt 10% nominal.

bandpass amplifier formulas:

select C3 and C4 to be 0.01 $\mu F$, then

$$R7 = \frac{Q}{f_O C3} \qquad R5 = \frac{R6}{2A} \qquad R6 = \frac{R5R7}{4Q^2R5 - R7}$$

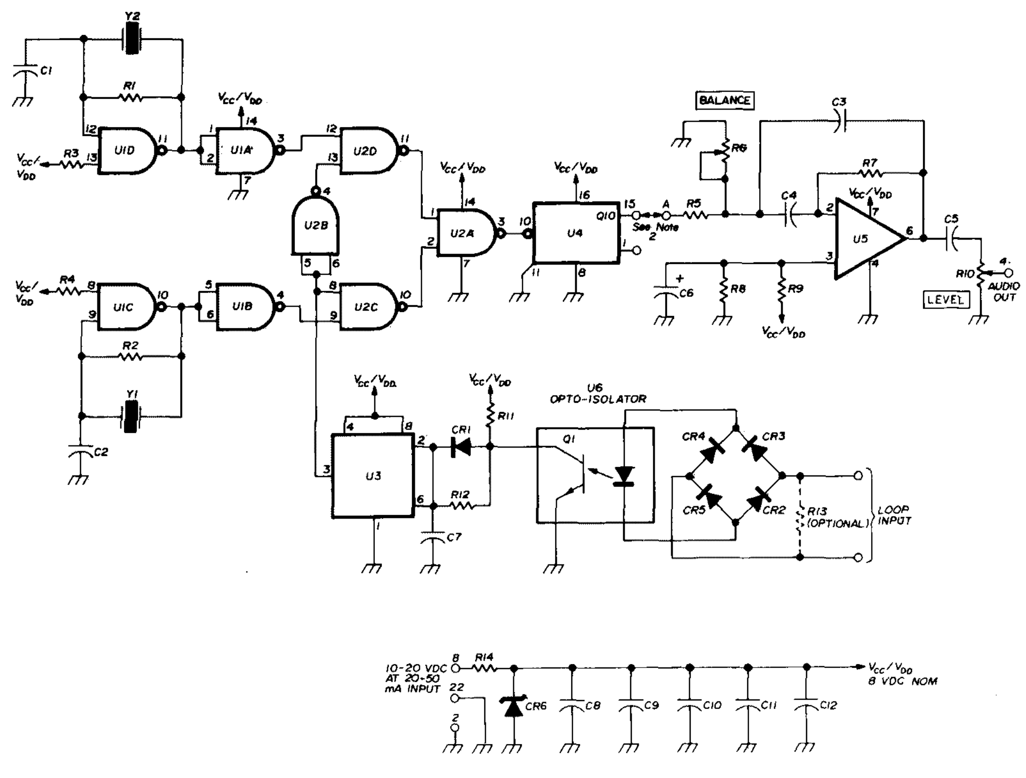
where Q = 10, A = desired gain, and $f_o$ = desired center frequency. The value of R6 should be chosen to allow variation over the desired range; e.g., if R6 is calculated at, say, 300 ohms for a given center frequency, then a 500-ohm pot should be used.

frequencies. R6 is adjusted to slew the filter center frequency, thus balancing the audio level of the mark and space tones. R10 is the audio output level adjustment. C3 and C4 are the only critical components in the system and should be stable with respect to temperature over the range anticipated in use. Mylar or polystyrene types are suggested, but disc ceramics may be used if environmental conditions are stable.

Note that U5 is operated single ended; that is, with only one power-supply polarity. This is accomplished by biasing the noninverting input at 0.5 $V_{cc}/V_{dd}$ with the network composed of R8, R9 and C6. C6 ensures a noise-free reference point. Because of the low-gain, moderate-Q characteristics incorporated in the filter, ringing is minimized and distortion to the switching rate is undetectable with the TU used for measurements.

Because of the imperfection of many surplus keyboards, debounce was incorporated in U3. Optical isolation with bridge input of the loop to the afsk was also incorporated to eliminate noise coupling and polarity sensitivity.

The operation of U3 and its components is as




NOTES: 1. C9-C12 ARE DISTRIBUTED $V_{CC}/V_{DD}$ BYPASSES.
2. JUMPER POSITION DETERMINED BY DESIRED OUTPUT FREQUENCY-SEE TEXT AND TABLE 1.
3. CONNECTOR PINOUT CONFORMS TO DT-600 WHERE PRACTICAL.
4. D/AFSK-100 BOARD COMPRISES TWO COMPLETE AFSKS. PINOUT ON ICs IS THE SAME FOR EACH AFSK.
5. PINOUT FOR SECOND AFSK: 12, 13 LOOP INPUT 19 AUDIO OUTPUT

**fig. 1. Schematic of the improved afsk design. Minimum parts and maximum circuit flexibility are featured.**

follows: Assuming the loop is closed, isolator transistor Q1 is saturated, and C7 is discharged through R12. When the loop opens, C7 charges through CR1 and R11. Since CR1 and Q1 have similar junction characteristics, charge and discharge times will be essentially identical with no contact bounce. If bounce on loop closure occurs discharge time will extend; if it occurs on opening charge time will extend.

U3 is connected as a Schmidt trigger whose trigger point is $1/3V_{cc}/V_{dd}$ and whose reset point is $2/3V_{cc}/V_{dd}$, thus providing symmetrical action with wide hysteresis. If debounce is not desired or needed, R12, CR1 and C7 may be deleted, with a jumper added in place of CR1. U3 is retained to insure fast mark-space transition. The loop input is designed for 20 mA operation. If 60-mA loop operation is desired, add a 33-ohm resistor in parallel with the isolator input (optional resistor R13 in fig. 1).

**table 2. Parts list. A ready-to-use PC board with assembly and troubleshooting information is available (see text).**

| | |
|---|---|
| Y1, Y2 | EX grade, 32 pF series resonant. See table 1 |
| U1, U2 | CD4011AE or equivalent cmos quad 2-input NAND gate |
| U3 | NE555V timer or equivalent |
| U4 | CD4020AE or equivalent cmos 14-stage ripple counter. |
| U5 | LM741CN or equivalent 741-type op-amp |
| U6 | H-11A or equivalent opto-isolator |
| R1, R2 | 1 meg ¼ watt 10% |
| R3, R4 R11, R12 | 10k ¼ watt 10% |
| R5 | see table 1 |
| R6, R10 | 1k trimpot CTS X201-R102B or equivalent. See also table 1 |
| R7 | see table 1 |
| R8, R9 | 2.2k ¼ watt 10% |
| R13 | 33 ohm ¼ watt 10% |
| C1, C2 | 47 pF ceramic disc |
| C3, C4 | 0.01µF mylar or polystyrene |
| C6 | 10 µF 15 wVdc aluminum electrolytic |
| C5 | 0.22 µF ceramic disc |
| C10-C12 | 0.1 µF ceramic disc |
| C8, C9 | 50 µF 15 wVdc aluminum electrolytic |
| C7 | 0.01 µF ceramic (60 wpm Baudot) or 0.005 µF ceramic (100 wpm ASCII) |
| CR1 | 1N914 or equivalent silicon diode |
| CR2-CR5 | 1N4002 or equivalent |
| CR6 | 1N959 zener, 8.2 volt, 400 mW or equivalent |

*An etched, drilled and plated PC board is available from Precision Technical Labs, P.O. Box 6743, Richmond, Virginia, 23230. A copy of this article plus component layout and troubleshooting information will be included with the board. The boards are laid out to accept two complete afsk circuits to allow simultaneous independent Baudot and ASCII operation. Order part number D/AFSK-100 and remit $10.95 (check or money order) for each board desired. Allow 2 to 3 weeks for delivery. Clubs and groups should write for quantity discounts. Boards are designed to conform to the DT-600 supply bus pinouts.

Power supply requirements are relaxed by including the zener regulator circuit R14, CR6, and C8-C12. This circuit allows operation from supply voltages between 10 - 20 Vdc. Current drain is 20 - 50 mA depending on supply voltage.

If operating frequencies other than the standard 2125-Hz mark and 2295-Hz space pair are desired, changes to the active filter and crystal frequencies may be calculated by the formulas given in **table 1** and in reference 6. The circuit is not designed for operation with shifts beyond 200 Hz since wide-shift RTTY is fast disappearing.

## construction and adjustment

Printed-circuit construction and good mechanical practices should be used in assembling this unit.* A parts list is provided in **table 2**. Since moderate rf levels are used, attention should be paid to parts placement. Setup is simple and should take little time after construction. Upon applying supply voltage to the unit, some level of tone should appear at the output. Set the output level control for maximum, and adjust the balance control for maximum output. The open-circuit output at this point should be about 1 to 1.5 Vac and should be the space tone. Connect the unit into the loop, and the tone should shift to the mark frequency. If this is the case, run a test tape and adjust the balance control for unchanging output amplitude. Adjust the output level control for the desired audio level, and the unit is operational.

## conclusion

This afsk unit was designed for optimum quality and stability in keeping with modern design methods and components. Ease of assembly, checkout, and troubleshooting is ensured by a minimum of adjustments and critical circuits. Most sources of noise, distortion and poor stability have been eliminated. By careful selection (or elimination, where practicable) of debounce time constants, this unit may be used at maximum speeds encountered in normal RTTY and data operation using an afsk. Total cost is under $30, including the PC board (less with a well-stocked junk box). Excluding the crystals, all parts should be available at most parts houses.

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**ham radio**

# reducing intermodulation distortion in high-frequency receivers

## Understanding and improving intermodulation performance of amateur band receivers

Judging by some recent articles in *ham radio*, amateurs are *finally discovering* that even the best of our high-frequency receivers are marginal when it comes to frontend performance in the presence of strong signals. Interestingly enough, the problem has not been caused by any lack of suitable technology, but rather most amateurs have simply been unable to clearly recognize and define the symptoms. We also have placed little pressure on manufacturers to produce receivers with cleaner frontends. Over the last several years, however, we have had to operate our receivers in increasingly dense rf environments. As a result, most of us have become well acquainted with a wide variety of frontend maladies in our communications receivers.

Recently, we have read about amateurs who have built receiver frontends employing special components and techniques that result in superior performance.[1,2] It is probably only a matter of time before the amateur receiver manufacturers incorporate some of these features into a new generation of amateur receivers. To a limited extent, this has already begun. Realistically speaking, I don't believe that most amateurs are contemplating homebrewing a receiver from scratch or unloading a present receiver for a new one in the near future. This then leaves two choices; live with the problem or modify our existing receivers to improve frontend performance. The first choice is easy, but I endeavored to determine if the second could reasonably be accomplished by a technically competent amateur. I ascertained that it can be done, but only with considerable time and effort.

### what's right about our present receivers

Before we tear apart the designs of existing receiver frontends, let's attempt to understand the philosophy behind those designs. You may suspect a concept that has lasted all these years can't be entirely bad, and would have some compensating virtues. This is certainly the case. **Fig. 1** is a simplified block diagram of a typical amateur receiver frontend. Dual conversion is used, with the first i-f high enough to obtain good image rejection and the second i-f low enough to permit high gain while providing the required selectivity. The first LO (local oscillator) is a fixed-frequency, crystal-controlled oscillator for high stability, while the second LO (the vfo) can be operated at a low frequency where it also can be made stable. Another factor that enhances vfo stability is the absence of a vfo bandswitching requirement. Bandswitching is accomplished by changing crystals in the first LO. What we really have here is a low frequency tunable i-f preceded by heterodyne converters. With this one basic design, we buy sensitivity, selectivity, image rejection, and frequency stability in one neat package. So, rest assured that there has been no conspiracy among the manufacturers to sell us receivers with inferior frontends. They have simply employed a design incorporating nearly all the frontend qualities that most amateurs have asked for in the past.

### what's wrong

Now let's flip the coin and take a hard look at the other side. In a receiver frontend, such as the one shown in fig. 1, we have three active stages preceding the second i-f filter. The composite bandwidth of the frontend could be on the order of hundreds of kHz, permitting strong in-band signals to find their way through all three stages. The second i-f filter eliminates all but the desired signal. Speaking in terms of frontend intermodulation, cross modulation, and desensitization, though, the damage has already been done. Simply stated, the active devices cannot handle the high signal levels produced as a result of the broadbanded gain. You might guess that the active stage most likely to cause trouble is the second mixer, since it is preceded by the most gain. Therefore, if you want to improve the frontend performance of a dual-conversion receiver similar to

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the one shown in fig. **1**, you should direct your efforts toward cleaning up the second mixer.

Before we discuss methods of cleaning up the second mixer, let's briefly review some of the more common receiver frontend maladies and what causes them. The problem that is probably most familiar to amateurs is receiver desensitization. This is caused by the presence of a strong signal close to the desired signal. Whenever the strong signal comes on the air, the desired signal amplitude decreases. Desensitization occurs because the strong signal has driven one or more of the active frontend stages into gain compression, thereby reducing the frontend gain for all signals.

Another frontend problem is cross modulation. This occurs when a strong signal, close in frequency to the desired signal, imparts some of its modulation to that signal. Again, the cause of cross modulation is nonlinearity in one or more of the active frontend stages.

When two signals at frequencies $f_1$ and $f_2$ interact in a nonlinear device, intermodulation (IM) products result. As long as these IM products fall outside the rf passband of the receiver frontend, no damage is done. Otherwise, there will be spurious responses as we tune the receiver across the band. IM products can be further categorized as being either even-order or odd-order. Even-order products (2nd, 4th, etc.) are normally of no concern in amateur receiver frontends, since either $f_1$ or $f_2$ (or both) would have to be well outside the receiver rt passband in order to produce in-band IM products. As long as the rf passband is less than an octave, as is ordinarily the case in an amateur receiver, even-order IM products will not be troublesome. Odd-order IM products (3rd, 5th, etc.) however, do cause trouble. They can be produced by in-band signals and no practical degree of frontend selectivity can eliminate them.

To summarize then, the three receiver frontend maladies that seem to give amateurs the most trouble are desensitization, cross modulation, and odd-order intermodulation products. Are separate remedies required? Fortunately, a remedy for any one of these problems is ordinarily applicable to the others. Therefore, we will only address the problem of odd-order IM distortion. More specifically, we will confine our efforts to the reduction of 3rd-order IM products, as higher odd-order products are usually less troublesome.

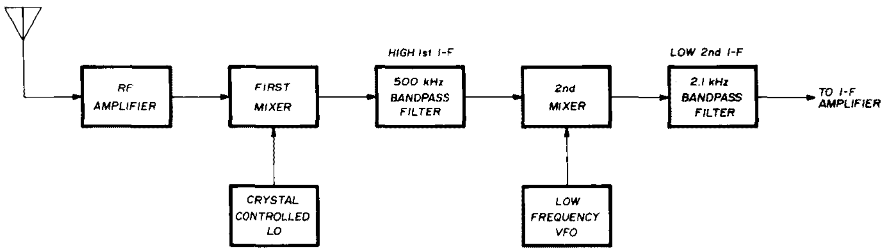



**fig. 1. Typical front-end of a dual-conversion amateur-band receiver.**

One of the reasons for the general lack of knowledge by amateurs concerning IM distortion has been the absence of an industry-wide method of specifying it. Among the few amateur receiver manufacturers who even bother to specify IM performance at all, no two receivers are specified in such a manner that one can be meaningfully compared against the other. The figures provided are often meaningless mumbo-jumbo anyway, since the measurement conditions are neither standardized nor specified. *What we really need then is a method of specifying IM performance that is universal and does not involve so much hand-waving.*

## intercept point method

A general and succinct manner of specifying the 3rd-order IM performance of a receiver (or any device) is the intercept-point method. Not only does the intercept-point method permit you to compare the IM performance of two different receivers, but it also actually lets you compute the level of the resultant IM products given the strength of the two offending in-band signals. The intercept point method is well discussed in a 1967 article by McVay.[3] It has also been discussed in recent amateur magazine articles.[1-4] Hopefully, the manufacturers will soon take the hint and adopt the 3rd-order intercept point as a standardized specification, as has been done elsewhere in the communications industry.

Since the 3rd-order intercept-point concept has been described elsewhere, let's not dwell on it too much here, except to define it and indicate its application. The general rule of thumb is that if we know the 3rd-order intercept point of a receiver frontend is $X$ dBm (0 dBm = 1 mW), and two in-band signals ($f_1$ and $f_2$) each have

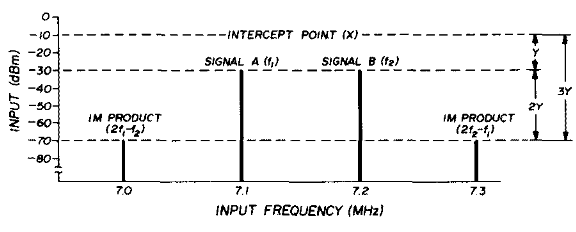



**fig. 2. Two signals and their resultant inband 3rd-order intermodulation products.**

an amplitude of $Y$ dBm below the intercept point, then the two resultant in-band 3rd-order IM products will have equivalent input amplitudes of $X - 3Y$ dBm. Note that everything is referenced to the receiver input.

Plugging in some real numbers, suppose that a receiver is known to have a 3rd-order intercept point of −10 dBm referenced to the input. Let's further assume that two signals within the receiver rf passband each have amplitudes of −30 dBm. $X$ is then equal to −10 dBm, and $Y$ = −10 dBm − (−30) dBm = 20 dB, and by plugging these values into $X - 3Y$ we get −10 − 3(20) = −70. Thus, the two resultant in-band 3rd-order IM products would each have equivalent input amplitudes of −70 dBm.

Fig. 2 is a graphical representation of this situation, similar to what you might observe on a spectrum analyzer connected to the second mixer output before the i-f filter. The frequency axis has been referenced to the actual incoming signal frequency rather than the intermediate frequency. In addition, the signal levels are all referenced to the receiver input. Signal A ($f_1$) and signal B ($f_2$) are the two −30 dBm signals and are on 7.1 and 7.2 MHz, respectively. Notice that the frequencies of the two resultant in-band IM products occur at $2f_1 - f_2$ (7.0 MHz) and $2f_2 - f_1$ (7.3 MHz). If the receiver was tuned to either 7.0 or 7.3 MHz, the IM product would be indistinguishable from another input of −70 dBm. Since a −70 dBm input level is better than S-9 on most receivers, you could expect interference to valid stations that we might be trying to copy on either frequency.

An interesting implication is that the IM products will increase in amplitude in proportion to the signals that cause them. Suppose that signals A and B increase in amplitude by 10 dB (to −20 dBm). Since the amplitude difference ($Y$) between these signals and the −10 dBm intercept point is now only 10 dB (−10 − (−20) = 10), the IM products will now have equivalent input amplitudes of $X - 3Y$ = −10 − 3(10) = −40 dBm. As a result, the amplitude of the IM products has increased a hefty 30 dB for an increase of only 10 dB in signals A and B. Fig. 3 illustrates this point. If we continue to raise the levels of signals A and B until they reach the −10 dBm intercept point, the IM products will also be −10 dBm ($X$ = −10; $Y$ = −10 − (−10) = 0; $X - 3Y$ = −10 − 0 = −10 dBm). If we were to simultaneously graph (fig. 4) the receiver output level of either signal A or B and either IM product output amplitude as a function of input signal amplitude, the lines would intercept each other at an input level of −10 dBm which, appropriately enough is called the "intercept" point. If, instead of raising the levels of signals A and B, we lower them, the level of the IM products will *decrease* at the same 1:3 ratio.

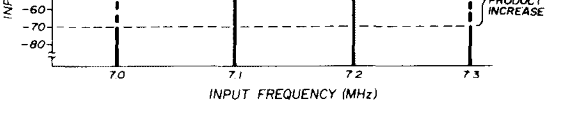



**fig. 3. The change in the IM product level as caused by a change in the fundamental signal levels. The IM products will increase by a factor of three as the signal increases. Conversely the IM products will decrease by the same factor.**

In the real world, the intercept point can only be inferred by extrapolation of the two lines, since the receiver front-end reaches its gain compression region at an input level that is typically 10 to 15 dB below the intercept point. As a result, the curves would tend to flatten out to the right and would never reach the intercept point.

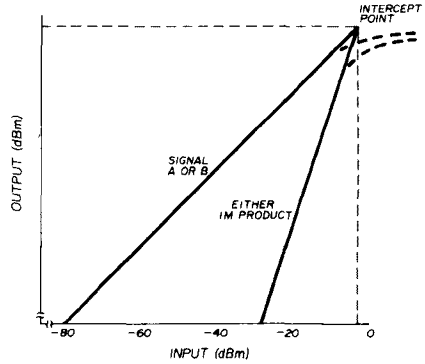



**fig. 4. Intercept point graph. The intercept point is really an extrapolation of the signal levels. The active stages will go into gain compression before the intercept is reached.**

The intercept point model presented above is valid as long as the receiver front-end is not being driven into its gain compression region. Also, we have assumed that signals A and B are equal in amplitude. If the two signals are different in amplitude, a correction factor must be applied. Finally, the rf bandpass characteristics of the receiver can modify the input levels of signals across an amateur band.

## receiver performance

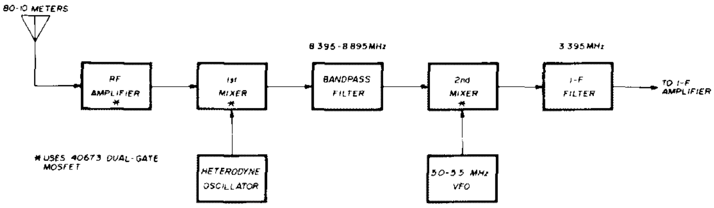
High-performance receivers exist today with 3rd-order intercept points in excess of +25 dBm, but this performance comes at a steep price. Fortunately, this type of performance is far better than most amateurs really need and can reasonably expect. The IM performance of commercially designed amateur receivers varies over a wide range. Interestingly enough, many of the new receivers on the market are more prone to IM distortion than the older ones. If you have a single-conversion, tube-type receiver, you are probably reasonably well off with regard to IM. On the other hand, the high sensitivity solid-state multi-conversion receivers tend to suffer more from IM distortion. In terms of round numbers, the 3rd-order intercept point of amateur receivers might be −40 dBm or even worse at the bottom of the scale, and perhaps −10 dBm or somewhat better at the high end. It's difficult to pinpoint what is "good" or "bad", but I feel that an amateur

receiver with a 3rd-order intercept point of −10 dBm or better will provide excellent IM performance in most cases.

## practical revisions

As mentioned earlier, in most dual-conversion receivers, the second mixer is the worst offender in terms of IM distortion, and any efforts to raise the 3rd-order intercept point should be directed toward improving the performance of this mixer. **Fig. 5** is a block diagram of my receiver front-end prior to modification. Note the similarity to **fig.** 1. Even though high quality fets are used throughout, severe IM distortion was experienced, particularly at night on 40 meters. Standing alone, the IM performance of the second mixer is not bad, but this performance is degraded by whatever gain precedes this mixer, which in this case is quite considerable. For example, if the second mixer had a 3rd-order intercept point of −10 dBm and was preceded by 30 dB of gain in the rf amplifier and first mixer, the overall receiver 3rd-order intercept point would be −40 dBm (assuming that the second mixer was the primary cause of the IM). A receiver with a 3rd-order intercept point of −40 dBm has *very* poor IM performance.

**fig. 5. Block diagram of the author's receiver.**




I decided that the best solution to the problem was to replace the fet second mixer with a hot-carrier diode double-balanced mixer. With proper vfo injection levels and correct termination, a mixer 3rd-order intercept point of +10 to +15 dBm can be realized. A Watkins-Johnson (relcom) M6 low-level, double-balanced mixer was chosen, although comparable units made by Anzac, Mini-Circuits Laboratory, and other firms could have been substituted with essentially the same results.

At the outset, it seemed that three basic problems had to be solved: building a low-noise, high-intercept point i-f post-amplifier to compensate for the 6 to 7 dB conversion loss of the M6, setting the vfo drive level so that a constant +7 dBm LO signal would be injected into the M6 mixer across the entire vfo tuning range, and changing circuit impedances to accommodate the 50-ohm port impedances of the M6. Unfortunately, solving these problems led to other problems that made the task considerably more difficult than I had first imagined.

The circuit that finally evolved is shown in **fig.** 6. Several points should be mentioned. For one thing, the M6 does not like reactive terminations at its i-f port, and to a lesser extent at the LO port. Reactive terminations at these two ports adversely affect the conversion loss and the IM performance. Thus, it is not permissible to terminate the i-f (dc-coupled) port with a filter, since it becomes highly reactive at all frequencies outside the passband. Consequently, a grounded-gate jfet amplifier, biased to present an input impedance of 50 ohms, immediately follows the M6. The U320 jfet is specifically designed to provide high transconductance and outstanding IM performance in grounded-gate applications (and accordingly these outstanding qualities are reflected in its price). Other suitable "super jfets" are the U310, CP643, and CP651. It is important that the jfet installed in this circuit has a 3rd-order intercept point no less than 6 to 7 dB below that of the M6. Otherwise, the jfet becomes the weak link in the chain and the full benefit of the superior IM qualities of the double-balanced mixer is not realized. The pi-network output circuit couples the 2000-ohm output of the jfet stage to the i-f filter (also 2000 ohms in this case).

Fortunately, sufficient vfo output was available to provide the required +7 dBm of drive at the M6 LO port. The 3 dB resistive pad on the LO port tends to swamp out any impedance variations. The loaded vfo maintains a reasonably constant output amplitude over its frequency range. The impedance matching problem was solved by using a low-Q impedance matching network to transform the high output impedance of the 8.395 - 8.895 MHz bandpass filter down to 50 ohms. Matching the 50-ohm output impedance of the M6 to the high-impedance i-f filter was accomplished using the jfet grounded-gate amplifier as explained earlier. There was no difficulty in matching the 25-ohm output impedance of the vfo to the M6 LO port.

The first iteration of the circuit did not use the vfo lowpass filter or the second 8.395 - 8.895 MHz (50 ohm) bandpass filter. In addition, no special precautions were taken to shield the added circuitry. The result was obnoxious birdies every few kHz across the entire receiver tuning range. This was very perplexing — the original idea was to *reduce* spurious responses.

After some investigation, I discovered that the birdies were caused by high-order harmonics of the vfo (generated in the double-balanced mixer) heterodyning with harmonics of the heterodyne oscillator located elsewhere in the receiver. Don't forget that double-balanced mixers are extremely broadband devices that do a very efficient job of mixing, even up to several hundred MHz.

Further investigation revealed that the harmonics of the heterodyne oscillator were entering the double-balanced mixer through all three ports. Thus, the entire

unit was placed in an aluminum mini-box and filtering was added at the LO and rf ports. The 50-ohm 8.395 - 8.895 MHz bandpass filter was needed (despite the existing 8.395 - 8.895 MHz bandpass filter) to increase rejection of first LO feedthrough via the first mixer.

Even after all these measures were taken, some birdies were still noticed. More investigation revealed still another problem – "contaminated" +15 volt lines. It turned out that the decoupling circuits in the +15 volt line were not adequate to prevent healthy jolts of the first LO energy getting everywhere in the receiver, including the double-balanced mixer. Therefore, I liberally applied a heavy dose of L-C decoupling circuits throughout the relevant sections of the receiver. The decoupling circuits consist of 0.001 $\mu$F bypass capacitors and small, single-layer rf chokes close-wound with no. 32 AWG (0.2mm) wire on 1 megohm, half-watt resistors. This involved considerable effort but it solved the problem.

The entire project from beginning to end was laborious and time consuming. Yet, the tremendous improvement in receiver IM performance justified the effort. I no longer notice IM distortion on 40 meters at night and strong local stations operating in the same band do not desensitize the receiver in all but the most extreme cases. An evaluation of the modified receiver front-end indicated that the second mixer was no longer the primary cause of IM distortion; that dubious distinction now belongs to the first mixer. There is really no reason why a repeat performance of the second mixer modification can't be made here, although the task could be somewhat more complicated. For one thing, the rf stage output circuitry would have to be modified to work into 50 ohms without degrading image rejection. In addition, the heterodyne oscillator signal would have to be amplified to a constant +7 dBm level for all amateur bands. I'm not saying that this all can't be done, but for the moment, I'll let someone else carry the ball.

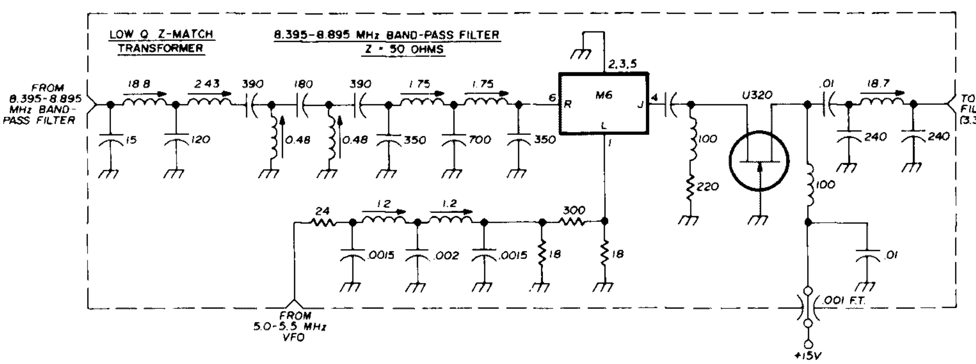



fig. 6. New second mixer incorporating the double-balanced mixer. All inductance values are in $\mu$H.

## other remedies

Using low-level, double-balanced mixers in a dual-conversion receiver substantially improves the frontend IM performance, but you do have other options. For example, the low-level mixers can be replaced with high-level types. These high-level double-balanced mixers require higher LO injection levels (up to one watt in some cases), but can provide 3rd-order intercept points in excess of +30 dBm, depending upon the device selected. For example, the Watkins - Johnson (Relcom) M9E yields a 3rd-order intercept point of typically +32 dBm with an LO injection level of +27 dBm. Active double-balanced mixers using jfets can also be built to provide similar performance, with the added advantage of providing conversion gain instead of the conversion loss inherent in passive double-balanced mixers. Unfortunately, high-level double-balanced mixers are quite expensive.

## conclusion

Not all amateur receivers are indentical, so I have intentionally been somewhat general about the modifications. Different receivers have different requirements, so if you want to modify your receiver, you will have to tailor the work to your specific situation. Therefore, the main point of this exercise has not been to provide a step-by-step method of cleaning up a particular receiver frontend, but to review the nature of the receiver frontend maladies and illustrate some of the considerations and problems involved in upgrading receiver frontend IM performance. My efforts in this regard have been successful and I'm sure that other amateurs who attempt to improve their receiver frontends can realistically expect similar or even better results.

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**ham radio**

# second-generation IC voltage regulators

## A survey of easy-to-apply second-generation ICs for use with inexpensive voltage-regulated power supplies

**Shortly after the monolithic** IC operational amplifier made its appearance as a low-cost circuit component, those intrepid linear IC designers who gave us the $\mu$A709 and similar op amps began plans to offer IC voltage regulators. This was not accidental, because the IC op amps had already found a major market in regulated power supplies. Early IC voltage regulators were relatively awkward to use, requiring a considerable number of external components to make them work. Fairchild had its $\mu$A723, National had its LM300, and Motorola had its MC1460. These ICs are still useful; and they, or their slightly improved descendants, are still found in new commercial designs (the LM300 has been superseded by the LM305, and the MC1460 by the MC1469).

In the last few years, however, a new generation of IC voltage regulators has appeared on the scene. These newer types are *much* easier to use than the earlier regulators, and some are so inexpensive that they have even cut into the zener diode market.

The National LM309 was one of the earliest three-terminal voltage regulator ICs to be introduced, and is typical of the new generation of such devices. It is a +5 volt (output) regulator designed to regulate $V_{cc}$ in TTL and DTL circuits for the immense digital logic market. The case of the LM309, normally grounded to the chassis for both electrical and heat conductivity, serves as the regulator's common lead and makes it really simple to use. As shown in fig. 1, the LM309 is available in two case styles: TO-3 and TO-39 (similar to TO-5), and supplies up to one ampere output current, depending upon the adequacy of the heatsink with which it is used.

By a curious quirk of fate, an *earlier* three-terminal voltage regulator made by Continental Devices Corporation (now a part of Teledyne), never really got off the ground. The CDC 513-4 was a +15 volt regulator enclosed in an inexpensive TO-5 plastic case. Had the CDC 513-4 been available with a different output voltage, or had it been released at a different time, perhaps it would have been the first successful three-terminal regulator. As it is, the CDC 513-4 is just an "also-ran" in the IC race.

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The success of the LM309 prompted further three-terminal regulator designs, and European Electronic Products (EEP) soon introduced the LM335, LM336, and LM337 in +5, +12 and +15 volts output versions, respectively; but available only in the TO-3 package.

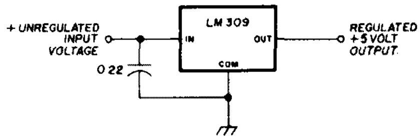



**fig. 1. Simple circuit illustrates extreme simplicity for using the National LM309 three-terminal IC voltage regulator. Note that case is grounded and that a capacitor is used between input terminal and ground.**

EEP is now known as Energy Electronic Products, and the foldback current-limited regulation curves of the EEP devices are shown in fig. 2.

The best attributes of the first three-terminal regulators of CDC, EEP, and National were combined in the Fairchild μA7800 series, which had the high-current capability of the LM309, the plastic package economy of the CDC 513-4, and the variety of output voltages of the EEP series. The Fairchild μA7805, 7806, 7808, 7812, 7815, 7818, and 7824 devices were packaged in either TO-3 enclosures or the less expensive TO-220 plastic power packages. The last two digits in each IC number identify the positive output voltage, and mean that +5, +6, +8, +12, +15, +18, and +24 volts, respectively, are available. Like the LM309 and EEP units, the case or metal-mounting tab is the common terminal, and is usually chassis grounded. Like the LM309 *and* the CDC 513-4, the μA7800 family is temperature-protected; that is, temperature rise of the chip controls shut-down. For this reason, the amount of current that the regulator will put out is dependent upon how well the heatsinking is accomplished, *and* which output voltage device you choose. For example, at 25°C the μA7805 will put out 500 mA, whereas the μA7824 will put out only 150 mA.

The Fairchild μA7800 family is perhaps the most popular, and also the most widely "second-sourced" three-terminal regulator of all. Similar devices, with similar part numbers, are produced by Motorola, National, Signetics, Silicon General, and others. These equivalents are shown in table 1.

## low- and medium-current regulators

The basic μA7800 family has been expanded to create the μA78M00 and the μA78L00 subfamilies; the difference being that these subfamilies, using smaller packages, are designed to limit at lower currents. If you remember the "M" is for medium current, and "L" is for low current, you'll get the general idea. The μA78M00 family is packaged in TO-220 and TO-39 enclosures, while the μA78L00 family is produced in TO-39 and TO-92 packages. The TO-92 enclosure is, of course, a small plastic inline transistor package. At 25°C the respective output currents of the μA7805, μA78M05, and μA78L05 are approximately 500 mA, 300 mA, and 40 mA. One other difference to be noted in the μA78L00 family is that it is available with an output of +2.6 volts, as well as the aforementioned higher values. As with the μA7800 family, the "M" and "L" series are second-sourced by other manufacturers. **Tables 2** and **3** cover the μA78M00 and μA78L00 equivalents.

## multivoltage regulators

National Semiconductor made a significant, if somewhat late, step forward with their LM340 and LM320 families of multivoltage, three-terminal IC voltage regulators. The LM340 series of devices is similar to the μA7800 series, and is produced in TO-3 and TO-220 packages. The companion LM320 series, however, is a family of ***negative*** voltage regulators, long needed by electronic circuit designers. The LM320 offered −5, −5.2, −6, −8, −12, −15, −18, and −24 volts. Now, the designer could easily construct a regulated power supply with outputs of ±5, ±6, ±8, ±12, ±15, ±18, and ±24 volts for operation of op amps and other linear ICs. Also, the important +12, −6 volt pair could be provided (for voltage comparators like the μA710 and μA711) as well as −5.2 volts for ECL ICs. Motorola and Fairchild quickly followed with negative-voltage three-terminal regulators offering the same output voltages, but with the addition of a −2 volt output member in their families. The three-terminal, negative-voltage regulators are shown in tables 4 and 5.

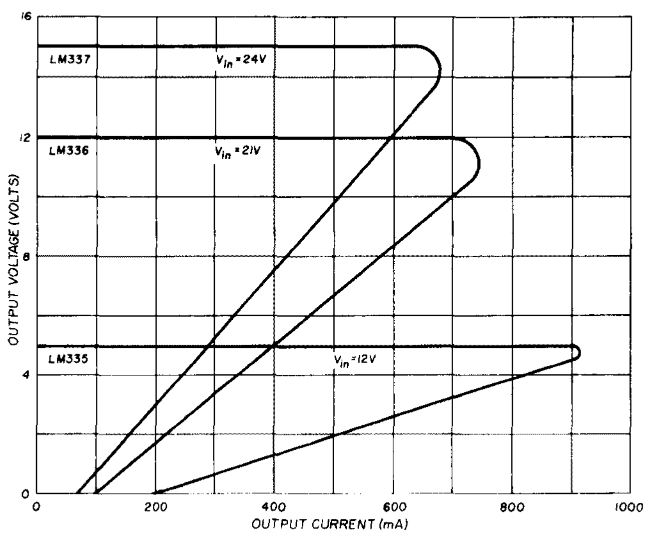



**fig. 2. Output voltage and current characteristics of EEP voltage regulators LM335, LM336, and LM337. Note the relationship between input and output voltage, and the current-limiting characteristics of each.**

A word of caution is appropriate when using these negative regulators: *do not ground the case*! The case is *not* the common terminal, but rather, the input terminal. This is unfortunate from the user's point of view, but apparently is dictated by the subtlties of IC chip design and a knowledge of which element must be at substrate potential. As with the positive voltage three-terminal regulators, an input capacitor is placed across the IC (and in close proximity, to provide a low inductance connection) for stability. Such a negative voltage regulator is shown in fig. 3.

In Europe, the SGS/ATES company produces the L129, L130, and L131 series of voltage regulators furnishing +5, +12, and +15 volt outputs. These are enclosed in TO-126 plastic power packages (the smaller of two *Thermo-pad* packages manufactured by Motorola in the U. S.). Electrically, the L129, L130, L131 regulators are remarkably similar to the EEP LM335, LM336, and LM337 types in that they also employ foldback current limiting, rather than thermal protection.

## high-current regulators

More recently, National introduced a higher current three-terminal regulator for +5 volt system regulation. The LM323 is a +5 volt, 3 amp device available only in a TO-3 case. Like its smaller relative, the LM309, the case is the common terminal, and – for stability – its input terminal requires a capacitor connected to the common terminal.

The National LM345 is similar to the LM323 in that it, too, is a 3-amp device, but provides negative output voltages. The LM345 is available with regulated outputs of −5.0 and −5.2 volts. As with the LM320 series, the case is the *input*, not the *common*, terminal. The LM345, also, is made only in a TO-3 package. National makes three other groups of proprietary three-terminal regulators designated as the LM341, LM342, and LM3910. The LM341 series is available with output voltages of +5, +6, +8, +12, +15, +18, and +24 volts, and is roughly comparable to the Fairchild μA78M00 family. The LM341 series is only available in a TO-202 plastic power package (similar to the package that General Electric uses for its plastic power transistors).

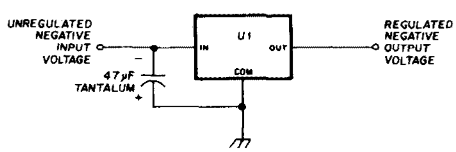



**fig. 3. Application of the three-terminal IC negative-voltage regulators such as the National LM320 or National μA7900 series. Note that the case must not be grounded. See text for details.**

The LM342 series is available in both TO-202 and TO-39 packages, and is also available with the same voltages as the LM341 family. The LM342 family has lower output current than the LM341 family.

A still-lower-current three-terminal IC regulator made by National is the LM3910 series packaged in TO-39 or TO-92 cases. This series is meant to replace the μA78L00 family with tighter output voltage tolerance, higher ripple rejection, better regulation, and lower

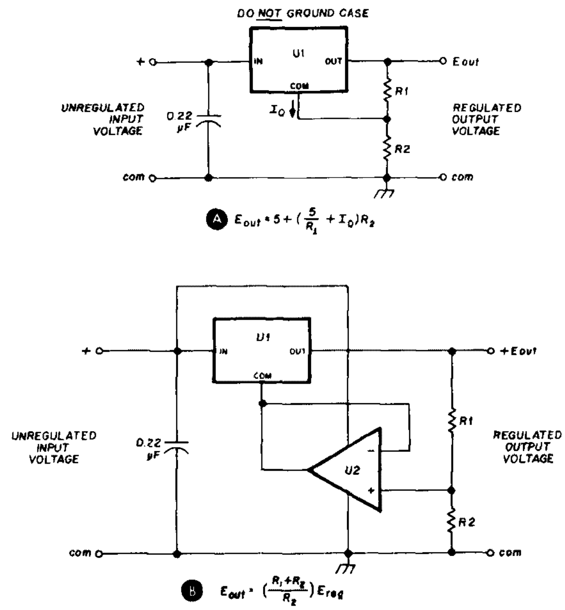



**fig. 4. The circuit in (A) shows a simple method of increasing nominal output voltage of a three-terminal IC voltage regulator such as the National LM340-05. R1, R2 form a simple voltage-divider network. The circuit in (B) shows how a three-terminal IC positive-voltage regulator can be combined with an op amp to increase the regulated output voltage. Text has details.**

quiescent current. To see the relative output currents of the LM341, LM342, and LM3910 families, let's compare the output current of each at +5 volts. At 25°C, the LM341-05 furnishes 500 mA, the LM342-05 furnishes 200 mA, and the LM3910-5.0 furnishes 100 mA.

## circuit applications

Many three-terminal IC regulators may be paralleled to provide higher current output. This is not advisable with the various LM309 types, but works well with the LM320, LM340, μA7800, and μA7900 ICs (and their equivalents).

There are several circuit tricks that may be played with three-terminal regulators to change their output voltages. The simplest one is shown in fig. 4A. Note that this technique can be used only to *raise* the regulated output voltage and that it slightly degrades regulation. The smaller the value of R2, the better the regulation, so it follows that the IC furnishing the closest voltage *on the low side* of the desired output voltage should be used.

A technique which allows variation of the output voltage uses an op amp as a non-inverting follower to

**table 1. Three-terminal positive IC voltage regulators by type number, voltage rating, and manufacturer. All are similar to the Fairchild μA7800 series. See text for current ratings and other characteristics. Blanks indicate that an equivalent IC is not available.**

| manufacturer | +5V | +6V | +8V | +10V | +12V | +15V | +18V | +20V | +24V | +28V |
|---|---|---|---|---|---|---|---|---|---|---|
| Fairchild | μA7805 | μA7806 | μA7808 | ----- | μA7812 | μA7815 | μA7818 | ----- | μA7824 | ----- |
| Lambda | LAS1505 | LAS1506 | LAS1508 | LAS1510 | LAS1512 | LAS1515 | LAS1518 | LAS1520 | LAS1524 | LAS1528 |
| Motorola | MC7805 | MC7806 | MC7808 | ----- | MC7812 | MC7815 | MC7818 | MC7820 | MC7824 | ----- |
| Motorola-HEP | C6110P | C6111P | C6112P | ----- | C6113P | C6114P | C6115P | ----- | C6116P | ----- |
| National | LM340-5 | LM340-6 | LM340-8 | ----- | LM340-12 | LM340-15 | LM340-18 | ----- | LM340-24 | ----- |
| National | LM7805 | LM7806 | LM7808 | ----- | LM7812 | LM7815 | LM7818 | ----- | LM7824 | ----- |
| Raytheon | RC7805 | RC7806 | RC7808 | ----- | RC7812 | RC7815 | RC7818 | ----- | RC7824 | ----- |
| Signetics | μA7805 | μA7806 | μA7808 | ----- | μA7812 | μ7815 | μ7818 | ----- | μA7824 | ----- |
| Silicon General | SG7805 | SG7806 | SG7808 | ----- | SG7812 | SG7815 | SG7818 | ----- | SG7824 | ----- |

**table 2. Medium-current (designated by the letter M), three-terminal, positive voltage-regulator IC by type number, voltage rating and manufacturer. See text for description.**

| manufacturer | +5V | +6V | +8V | +10V | +12V | +15V | +18V | +20V | +24V |
|---|---|---|---|---|---|---|---|---|---|
| Fairchild | μA78M05 | μA78M06 | μA78M08 | ----- | μA78M12 | μA78M15 | μA78M18 | μA78M20 | μA78M24 |
| Motorola | MC78M05 | MC78M06 | MC78M08 | ----- | MC78M12 | MC78M15 | MC78M18 | MC78M20 | MC78M24 |
| National | LM341-5 | LM341-6 | LM341-8 | ----- | LM341-12 | LM341-15 | LM341-18 | ----- | LM341-24 |
| Signetics | μA78M05 | μA78M06 | μA78M08 | ----- | μA78M12 | μA78M15 | μA78M18 | μA78M20 | μA78M24 |
| Motorola | MC7705 | MC7706 | MC7708 | ----- | MC7712 | MC7715 | MC7718 | MC7720 | MC7724 |
| Teledyne | 78M05 | 78M06 | 78M08 | ----- | 78M12 | 78M15 | ----- | 78M20 | 78M24 |

**table 3. Low-current (designated by the letter L), three-terminal, positive voltage-regulator ICs by type number, voltage rating, and manufacturer. These devices are very popular for regulating single op amp or linear IC stages.**

| manufacturer | +2.6V | +5V | +6V | +8V | +10V | +12V | +15V | +18V | +20V | +24V |
|---|---|---|---|---|---|---|---|---|---|---|
| Teledyne | ----- | ----- | ----- | ----- | ----- | 829 | 830 | ----- | ----- | ----- |
| Fairchild | μA78L02 | μA78L05 | μA78L06 | ----- | ----- | μA78L12 | μA78L15 | ----- | ----- | ----- |
| Motorola | ----- | MC78L05 | ----- | MC78L08 | ----- | MC78L12 | MC78L15 | MC78L18 | ----- | MC78L24 |
| National | ----- | LM342-5 | LM342-6 | LM342-8 | LM342-10 | LM342-12 | LM342-15 | LM342-18 | ----- | LM342-24 |
| National | ----- | LM3910-5 | LM3910-6 | LM3910-8 | LM3910-10 | LM3910-12 | LM3910-15 | LM3910-18 | ----- | LM3910-24 |
| National | ----- | LM78L05 | ----- | LM78L08 | ----- | LM78L12 | LM78L15 | LM78L18 | ----- | LM78L24 |
| Signetics | μA78L02 | μA78L05 | μA78L06 | ----- | ----- | μA78L12 | μA78L15 | ----- | ----- | ----- |
| Plessey | ----- | SL78L05 | SL78L06 | SL78L08 | ----- | SL78L12 | SL78L15 | SL78L18 | SL78L20 | SL78L24 |

**table 4. Three-terminal negative IC voltage regulators by type number, voltage rating, and manufacturer. All are similar to the Fairchild μA7900 series. See text for current ratings and other operating characteristics.**

| manufacturer | −2.0V | −5.0V | −5.2V | −6V | −8V | −12V | −15V | −18V | −24V |
|---|---|---|---|---|---|---|---|---|---|
| Fairchild | ----- | μA7905 | μA7952 | μA7906 | μA7908 | μA7912 | μA7915 | μA7918 | μA7924 |
| Motorola | C6117P | C6118P | C6119P | C6120P | C6121P | C6122P | C6123P | C6124P | C6125P |
| Motorola | MC7902 | MC7905 | MC7905.2 | MC7906 | MC7908 | MC7912 | MC7915 | MC7918 | MC7924 |
| National | ----- | LM320-5 | LM320-5.2 | LM320-6 | ----- | LM320-12 | LM320-15 | LM320-18 | LM320-24 |
| Silicon General | ----- | SG320-5 | SG320-5.2 | ----- | ----- | SG320-12 | SG320-15 | ----- | ----- |

**table 5. Low- and medium-current, three-terminal negative IC voltage regulators manufactured by Fairchild and Motorola. The low-current version is designated by a letter L in the part number, while the letter M indicates medium-current capability.**

| manufacturer | −3V | −5V | −5.2V | −6V | −8V | −12V | −15V | −18V | −20V | −24V |
|---|---|---|---|---|---|---|---|---|---|---|
| Motorola | MC79L03 | MC79L05 | ----- | ----- | ----- | MC79L12 | MC79L15 | MC79L18 | ----- | MC79L24 |
| Fairchild | ----- | μA79M05 | ----- | μA79M06 | μA79M08 | μA79M12 | μA79M15 | ----- | μA79M20 | μA79M24 |

sample the output voltage divider and drive the common terminal of the three-terminal regulator. This circuit overcomes the degradation of regulation of the simpler circuit shown in fig. 4A. The op amp variation is shown in fig. 4B. Again, only voltages higher than the nominal regulated voltage may be achieved.

The circuit of fig. 5 shows a "real-value" circuit which is capable of output voltages from +7 to +20 volts. Reference 1 shows many more uses for three-terminal IC voltage regulators, but many are so complex as to defeat the purpose of these simple ICs.

Three-terminal voltage regulators have made it possible for commercial firms to offer simple, inexpensive modular power supplies that even experimenters can afford. For instance, for $8.95, plus $1.25 handling, Adva* offers kits including PC board, transformers, and

*Adva Electronics, Box 4181 BE, Woodside, California 94062.

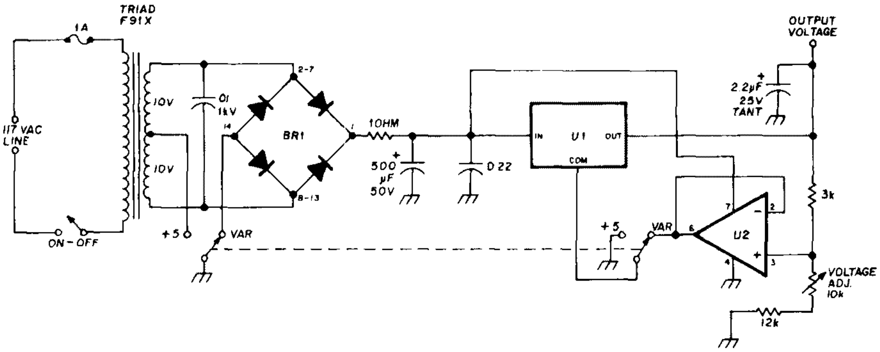
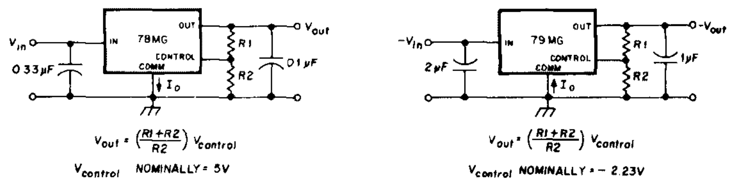



**fig. 5. Circuit capable of providing fixed-variable output voltages; for example, +5V at 200 mA or 7-20V at 100 mA. Heatsink tab of U1 must be connected to floating heatsink. BR 1 is Adva bridge, U1 is LM340-05, and U2 is National LM741 CM op amp.**

parts for construction of +5, +6, +9, +12, and +15 volt fixed, voltage-regulated power supplies. These supplies are based on the National LM340 series of three-terminal IC voltage regulators. Since all the Adva power supplies use the same type PC board, and the 9-volt unit uses two circuits like that of **fig.** 4A, output voltages are readily trimmable by adding a couple of resistors.

voltage for powering op amps, it is also the most common voltage rating for fixed, dual voltage regulators. There are also several dual positive-negative regulators available, such as the Silicon General SG3501, Motorola MC1468, and Raytheon RC4194, which may be used to produce *adjustable* dual outputs. These devices require one or more external resistors to establish the desired

**fig. 6. Fairchild 78MG and 79MG four-terminal IC voltage regulators. The 79MG can provide +5 to +30 volts and 79MG can provide −2.2 to −30 volts, regulated. Current through R2 should be approximately 1 mA; therefore, R2 = 5k ohms for the 78MG, and 2.2k ohms for the 79MG.**




Two recent ICs introduced by Fairchild are the 78MG and 79MG four-terminal voltage regulators. The 78MG is a positive voltage, and the 79MG a negative voltage regulator. They operate in the same general manner as the combination three-terminal voltage regulator and op amp of **fig. 4B**, requiring only a resistive divider to set the output voltage. Circuits using the 78MG and 79MG regulators are shown in fig. 6. These two ICs are offered in a plastic power package that has four pins on what appears to be a half-DIP package, with heatsink "ears" protruding from each side. The package is shown in fig.

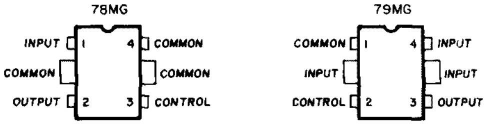



**fig. 7. Top views of package connection diagram for Fairchild 78MG and 79MG voltage regulators of fig. 6.**

**7.** The 78MG and 79MG regulators may be set for output voltages from +5 to +30 and −2.2 to −30 volts.

Some regulators are devoted to fixed *dual* output voltages. Inasmuch as ±15 volts is the most common

output voltages. Dual regulators built around these units are shown in **fig. 8**. The current capability of all three of these ICs may be increased by adding external power transistors and other discrete components, as discussed in references 2, 3, and 4.

Raytheon's RC4195 and National's LM325, 326, and 327, however, take the dual-voltage regulator a step further in the direction that made the three-terminal regulator so popular. These four ICs offer *fixed*, dual output voltage and simplicity of application — as well as an inexpensive power package option. The Raytheon RC4195 is available in a 9-pin TO-66, an 8-pin TO-5 and a plastic half-DIP. The TO-66 (RC4195TK) has the greatest heat dissipation, and may be used up to currents of 100 mA. The smaller RC4195T and RC4195DN (TO-5 and half-DIP) are better suited for spot-regulating op amps in which lower currents are required. A simple ±15 volt regulator using the RC4195T or RC4195DN is shown in **fig. 9**. Note that the pin connections would be changed if the RC4195TK unit were used.[4]

The National LM325, LM326, and LM327 are for ±15, ±12, and +5, −12 volts, respectively. They are available in the 14-pin DIP (suffix N), the 14-pin DIP power package (suffix S), and the 10-pin TO-100 package (suffix H). The TO-100 package is similar to the

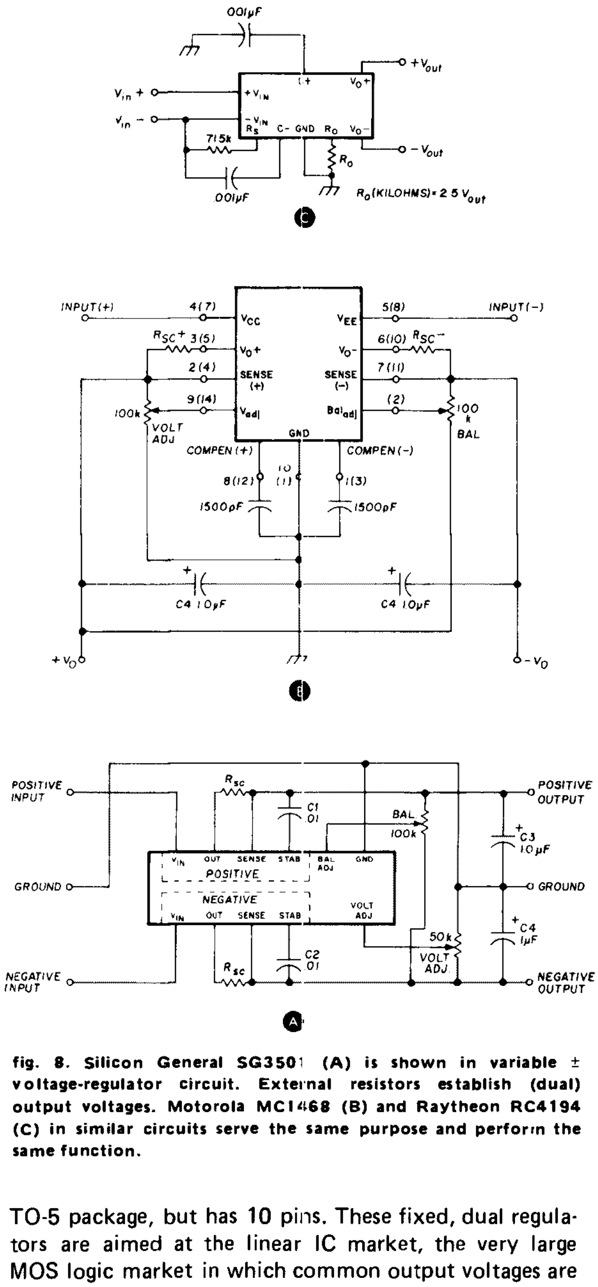



**fig. 8. Silicon General SG3501 (A) is shown in variable ± voltage-regulator circuit. External resistors establish (dual) output voltages. Motorola MC1468 (B) and Raytheon RC4194 (C) in similar circuits serve the same purpose and perform the same function.**

TO-5 package, but has 10 pins. These fixed, dual regulators are aimed at the linear IC market, the very large MOS logic market in which common output voltages are +5, and −12 volts.

The S package is similar to that used by SGS/ATES

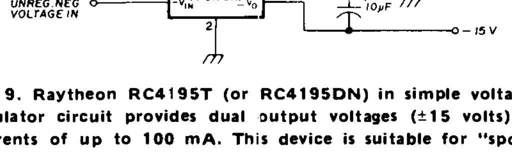



**fig. 9. Raytheon RC4195T (or RC4195DN) in simple voltage-regulator circuit provides dual output voltages (±15 volts) at currents of up to 100 mA. This device is suitable for "spot" regulation of op amps.**

and others in their audio power amplifier ICs in that it has a thick metal strap running the length of the DIP, which is bent over at the ends. The idea is to solder these bent-over ends into the copper lands of a PC board, to facilitate heat dissipation. The use of these ICs as dual regulators is shown in **fig. 10**. Different pin connections will be required if the N or H packages are used. With good heatsinking, the S version is capable of up to 100 mA output. The LM325, 326, 327 family may also be used with external power transistors to increase their current capability; this is covered in reference 1.

## higher-current regulators

Finally we come to the brutes of the voltage regulator scene, the MPC900 (negative) and the MPC1000 (positive). These are *not* fixed-voltage regulators at all, but they do represent the ultimate (to date) in current capability for single-regulator ICs. Ten amperes is the maximum output current of these devices, and they may be set to provide output voltages of −4 to −30 volts and +2 to +35 volts, respectively. **Fig. 11** shows these ICs in useful circuits. Of course, very low thermal resistance heatsinking should be used with these ICs if maximum current is to be drawn from them. These ICs are relatively expensive, and most amateurs will probably choose smaller and more complicated, but less expensive IC voltage regulators with external power transistor(s) to boost output current.

## summary

In this article I have attempted to cover only the newer generation of IC voltage regulators, touching only briefly, for comparison purposes, on the older types. The older types are still useful, and many companies still elect to use them in new designs because they are so widely second-sourced and have become so inexpensive. References to new ICs have, in this article, been made to the least expensive commercial grades of each IC, which is what amateurs find most available. It should also be borne in mind that all of these IC types are *series* regulators, and require a certain minimum difference voltage across the chip — typically between 2 and 3 volts. The input (unregulated) voltage *must* be at least 2 or 3 volts higher than the output (regulated) voltage.

Most regulator ICs require some sort of external compensating capacitor, and will oscillate without this

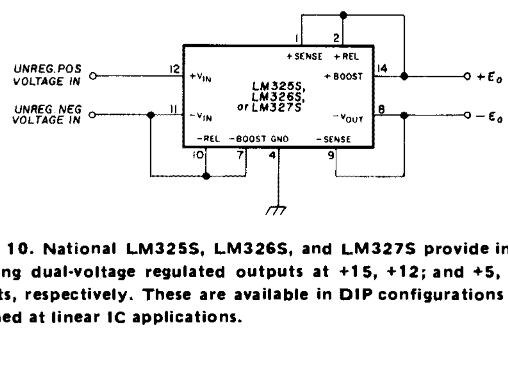
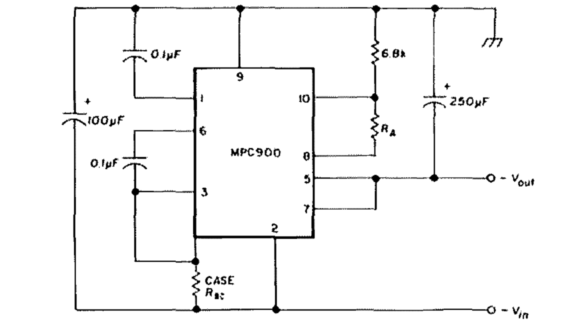
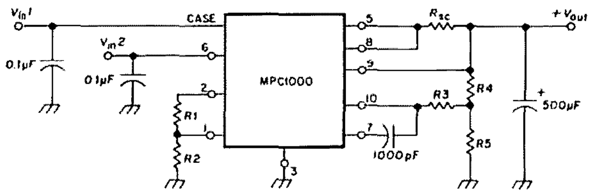



**fig. 10. National LM325S, LM326S, and LM327S provide interesting dual-voltage regulated outputs at +15, +12; and +5, −12 volts, respectively. These are available in DIP configurations and aimed at linear IC applications.**




B




A

PARAMETER VALUES FOR BEST RESULTS

| | $2V < V_O < 7V$ | $7V < V_O < 35V$ |
|---|---|---|
| R1 | $\approx \frac{R2(7-V_O)}{V_O}$ | $= \left[\frac{R4R5}{R4+R5}\right]$ |
| R2 | $10k < R1 + R2 < 100k$ | $\infty$ |
| R3 | $= \left[\frac{R1R2}{R1+R2}\right]$ | – |
| R4 | – | $\approx \frac{R5(V_O-7)}{7}$ |
| R5 | $\infty$ | $10k < R5 < 100k$ |
| $R_{sc}$ | $\approx \frac{.066}{I_{sc}}$ @ $T_J = 25°C$ | |

**fig. 11. High-current, up to 10 amperes, can be handled by the MPC1000 and MPC900 IC devices; the MPC1000 is for positive output voltages and the MPC900 is for negative output voltages. These regulators may be set from +2 to +35 and −4 to −30 volts, but require special heatsinking.**

capacitor. This capacitor *must* be wired close to the IC to minimize the inductance of the leads. It is always a good idea to check the manufacturer's data sheet of the regulator IC for the connections of each particular package because different package versions of the same IC can have quite different connections. Finally, the same spec sheet should also be consulted to find out what the IC's built-in heatsink is connected to. Arbitrary grounding of the heatsink tab or case can spell instant demise of the IC.

## references

1. *Voltage Regulator Handbook*, section 7, National Semiconductor Corporation, Santa Clara, California, 1975.
2. Silicon General Technical Bulletin 1501, "Dual Voltage Tracking Regulator," October, 1970.
3. Motorola Semiconductor DS9213, "Dual ±15 Volt Regulator," October, 1972.
4. *Total Linears Handbook*, Raytheon Company, Semiconductor Division, Mountain View, California.

**ham radio**

# the polarization diplexer — a polaplexer

## The polaplexer combines orthogonally-polarized signals from receiver and transmitter to permit full duplex operation

The polaplexer is a coined name for the POLArization-diPLEXER used by the San Bernardino Microwave Society (SBMS) to provide isolation between a microwave transmitter and receiver. This method uses the transmitter as the receiver local oscillator, thereby economizing on microwave sources. The polaplexer allows duplex communications (simultaneous transmission and reception) with minimum power loss. The use of the polaplexer within the complete communications system has been covered in past articles.[1,2,3]

The polaplexer uses circular waveguide as a transmission line, supporting orthogonal (at right angles) polarization for transmitting and receiving. In use, the polarizations are neither vertical nor horizontal, but at angles of 45 degrees to the vertical. The standard in use by the SBMS is that the polarization of the transmitted wave from all antennas is 45 degrees to the right, looking in the direction of transmission; the polarization of the received signal, orthogonal or at right angles to the transmitted signal, is 45 degrees to the left, looking in the direction of the other station. It can be seen in **fig. 1** that at the receiving site the signal coming from the distant transmitter will be polarized properly to be intercepted.

### waveguides and polaplexers

The circular waveguide used for construction of the polaplexer does not have to be silvered brass. Ordinary tin cans can be used if they have the proper diameter. The circular waveguide can be used as a feedhorn for a parabolic reflector or used to excite a circular horn. As will be shown, the circular waveguide must be within a proper size range for the frequency being used. The polaplexer operates in the dominant circular waveguide mode known as the $TE_{11}$ mode; the field configuration is shown in **fig. 2**. In order not to excite other waveguide modes, the waveguide diameter must be selected to exclude other modes. The waveguide size must be such that it will sustain propagation above the cutoff frequency of the $TE_{11}$ waveguide mode and below the cutoff frequency of the next higher $TM_{01}$ mode. The cutoff frequency is that frequency below which the attenuation of the waveguide, to that particular mode, increases at a rate which makes the waveguide useless except as an attenuator. To support the desirable $TE_{11}$ mode in circular waveguide the cutoff frequency, $f_c$, is defined by

$$f_c(TE_{11}) = \frac{175698.51}{d\ (mm)} = \frac{6917.26}{d\ (inches)} \qquad (1)$$

where

$f_c$ = cutoff frequency in MHz for $TE_{11}$ mode
$d$ = waveguide inside diameter

The $TM_{01}$ mode can be supported in circular waveguide whose cutoff frequency is defined by

$$f_c(TM_{01}) = \frac{229485.13}{d\ (mm)} = \frac{9034.85}{d\ (inches)} \qquad (2)$$

Another term that will be used in calculations is guide wavelength, $\lambda g$. The wavelength in a circular waveguide always exceeds the free-space wavelength and is related

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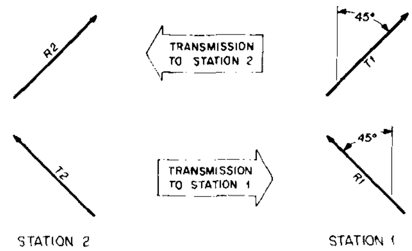


**fig. 1. Polarization standards adopted by the San Bernardino Microwave Society.**

to the cutoff frequency and operating frequency by the following equation

$$\lambda g = \frac{299792.5(mm)}{\sqrt{f_o{}^2 - f_c{}^2}} = \frac{11802.85(inches)}{\sqrt{f_o{}^2 - f_c{}^2}} \qquad (3)$$

where

$\lambda g$ = guide wavelength
$f_o$ = operating frequency
$f_c$ = $TE_{11}$ waveguide cutoff frequency

In practice two operating frequencies are used, separated by the i-f frequency. The operating frequency should be the geometric mean of the two operating frequencies or

$$f_o = \sqrt{f_{01} f_{02}} \qquad (4)$$

where

$f_o$ = mean operating frequency
$f_{01}$ = lower communications frequency
$f_{02}$ = upper communications frequency

Substituting the second equation into the first,

$$\lambda g = \frac{299792.5(mm)}{\sqrt{f_{01} f_{02} - f_c{}^2}} = \frac{11802.85(inches)}{\sqrt{f_{01} f_{02} - f_c{}^2}} \qquad (5)$$

## probes

Since many of our microwave devices have coaxial terminations, we must have a transition from coaxial (TEM mode) to the circular waveguide $TE_{11}$ mode. This can be accomplished by inserting a 1/4-wavelength ground plane or probe. It should be placed 1/4-guide wavelength from the closed end of the waveguide as shown in **fig. 3**.

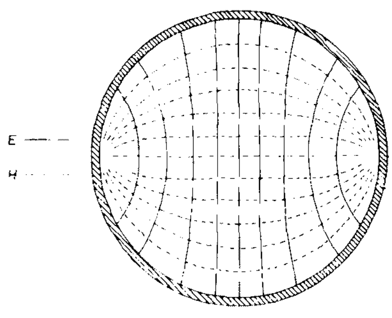


**fig. 2. $TE_{11}$ field configuration in circular waveguides.**

In the polaplexer, the transmitter and receiver terminations are orthogonal. If they were both placed 1/4-guide wavelength from the closed end of the polaplexer, they would touch. Further, even if they had no physical contact, the close proximity would reduce the desired isolation between them. Fortunately, in a transmission line impedances repeat each ½ wavelength along the line. If the orthogonal coaxial-to-waveguide transition is placed ½-guide wavelength toward the open end of the circular waveguide from the other transition, both polarizations can be excited in the waveguide with 20 to 40 dB isolation between them. **Fig. 4** depicts the orthogonal transition.

To make the basic polaplexer complete, a method of controlled injection of the transmitter signal into the receiver input must be devised because the transmitter is used as the receiver local oscillator. The first thought might be to place an injection screw halfway between the two orthogonal probes. Wrong! Well, half wrong anyway. Halfway between the two probes is ½-guide wavelengths from the closed end of the waveguide. The injection screw will do little more at this location than isolate the inboard probe. Therefore, the injection screw

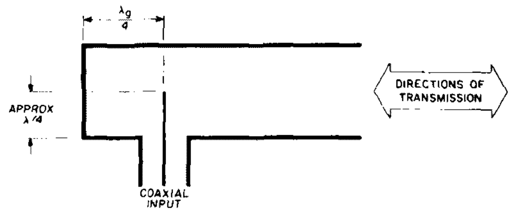


**fig. 3. Circular waveguide to coaxial transition.**

should be placed 1/3-wavelength (spacing arbitrary on my part) from the closed end of the waveguide at an angle which bisects the orthogonal probe polarizations. The designer has a choice of equally effective locations to select from.

## polaplexer design

The total length of the polaplexer should approach an odd number of quarter-guide wavelengths. This will provide the best impedance match within the waveguide. A waveguide that is not perfectly matched becomes a form of a resonant transmission line. If you assume you have a resonant transmission line and design the polaplexer length with this in mind, you cannot go wrong, even if you operate with a 1.00:1 vswr.

It will be shown that, using the three equations presented previously, a can with an inside diameter of 2-9/16 inches (about 2.56 inches or 65.0 mm) can be used as a polaplexer for the 3300-MHz amateur microwave band. First, the cutoff frequency for the $TE_{11}$ mode is

$$f_c = \frac{175698.51}{65.0} = 2699.4 \text{ MHz}$$

That tells us the waveguide will support the $TE_{11}$

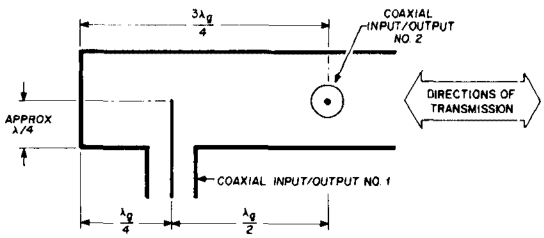


**fig. 4. Circular waveguide with orthogonal coaxial inputs.**

mode above 2700 MHz. Second, the cutoff frequency for the $TM_{01}$ mode is

$$f_c = \frac{229485.13}{65.0} = 3525.8 \text{ MHz}$$

This tells us that if we operate below 3525 MHz we will not excite the $TM_{01}$ waveguide mode. The third important parameter is guide wavelength. The 3300-MHz operating frequencies in southern California are 3335 MHz and 3365 MHz. Applying these and the $TE_{11}$ mode cutoff frequency to eq. 5 we obtain

$$\lambda g = \frac{299792.5}{\sqrt{3335 \cdot 3365 - (2699.42)^2}} = 151.1 mm$$
$$= 5.95 \text{ inches}$$

The dimensions needed to design a polaplexer are

| | | | | |
|---|---|---|---|---|
| 1/4λg | = | 37.8 mm | = | 1.49 inches |
| 3/4 λg | = | 113.3 mm | = | 4.46 inches |
| 5/4 λg | = | 188.9 mm | = | 7.44 inches |
| 7/4 λg | = | 264.5 mm | = | 10.41 inches |
| 1/3 λg | = | 50.4 mm | = | 1.98 inches |

Since the 7/4-λg length is a bit long, the 5/4-λg polaplexer length should be selected as a goal. The actual construction of a polaplexer is usually custom designed for and by the individual dependent upon the designer's resources. One configuration that I have used, on both the 3300 and 5700-MHz amateur bands, is illustrated in fig. 5.

One problem facing the polaplexer designer who uses simple probe injection is the probe length. An empirical equation that I have used successfully is

$$\ell = \frac{74948.13\,(1 + P - \sqrt{P})}{f_o} \text{ (mm)}$$

$$\ell = \frac{2950.71\,(1 + P - \sqrt{P})}{f_o} \text{ (inches)} \qquad (6)$$

where

$\ell$ = probe length
$f_o$ = mean operating frequency
$P$ = periphery of probe
$= \frac{\pi D}{\lambda_o}$ for circular cross section
$D$ = probe diameter

For example, assume the probe to be a no. 18 AWG (1mm) copper wire and the operating band to be 3300 MHz. The wire diameter is 1.0 mm (0.0403 in.). The operating wavelength is

$$\lambda_o = \frac{299792.5}{f_o} \text{ (mm)} = \frac{11802.85}{f_o} \text{ (inches)}$$
$$= \frac{299792.5}{\sqrt{3335 \cdot 3365}} = 89.5 \text{ mm} = 3.52 \text{ inches}$$

The periphery of the probe is

$$P = \frac{\pi D}{\lambda_o} = \frac{\pi 1.0}{89.5} = 0.035$$

The length therefore is

$$\ell = \frac{74948.13\,(1 + 0.0359 - \sqrt{0.0359}\,)}{\sqrt{3335 \cdot 3365}}$$
$$= 18.9357 \text{ mm} = 0.7455 \text{ inches}$$

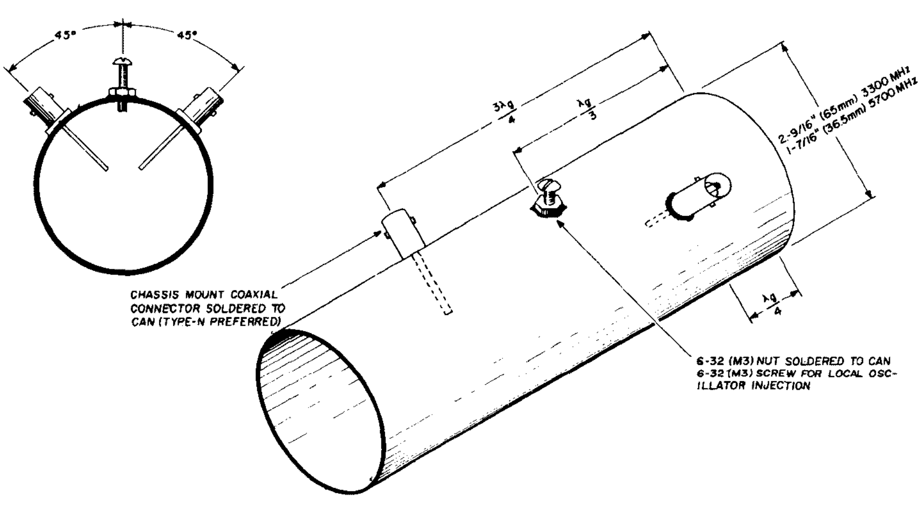


**fig. 5. Typical polaplexer showing layout and dimensions of principle parts.**

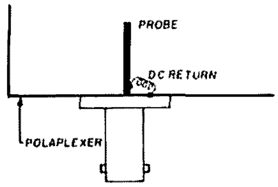


**fig. 7. Ground return for the receiver port in a polaplexer. The coil is 3 or 4 turns no. 30 AWG (0.25mm) wound over a piece of no. 20 AWG (0.8mm) wire as a coil form. The coil is soldered to the probe and the side of the polaplexer after removing the coil form.**

## mixers

There are many methods of providing a satisfactory transition from the coaxial TEM mode to the circular waveguide $TE_{11}$ mode. The method illustrated in fig. 5 is perhaps the simplest. However, it has the disadvantage of not providing the necessary dc return for a simple crystal mixer. I prefer using a coaxial-crystal mixer that provides the dc return with a tunable stub. Although the many and varied means of providing a dc return for a crystal mixer are beyond the scope of this effort, one method which I have used is shown in fig. 6. It provides a complete tunable mixer assembly. The BNC output subassembly was obtained from a surplus rectangular-waveguide mixer assembly and includes the necessary microwave-bypass capacitor to provide efficient conversion. The insulator in fig. 6 is the capacitor dielectric.

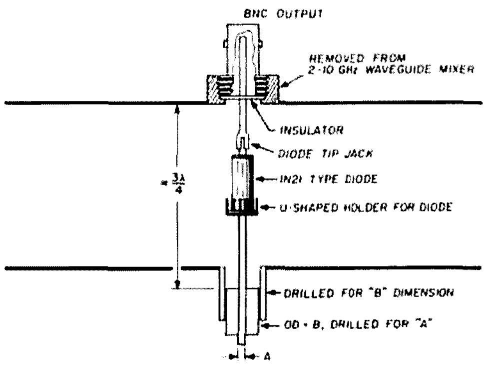


**fig. 6. Coaxial crystal mixer.**

The actual sizes of the adjustable short are left up to you and your own resources. The tolerances should be close. The center adjustment (identified by dimension B) could be threaded if desired and locking hardware used to maintain the adjustment once optimized. The crystal can be any good IN21-type crystal diode.

## references

1. Richard Kolbly, K6HIJ, "Standards for Amateur Microwave Communications," *ham radio*, September, 1969, page 54.
2. Richard Kolbly, K6HIJ, "A Low Cost Amateur Microwave Antenna," *ham radio*, November, 1969, page 52.
3. Richard Kolbly, K6HIJ, and Ed Munn, W6OYJ, "Microwave DX - California Style," *QST*, September, 1970, page 88.

ham radio

# 24-hour clock

## with digital readout and line-frequency time base

Complete construction details for a four-digit clock which uses a total of six ICs

**During recent** visits to other amateur stations, I noticed the increased use of electronic digital displays. These range from displays on very expensive transceivers, both hf and vhf, to frequency counters, clocks, voltmeters, thermometers, and even one that converts received CW signals to an alphanumeric display. Some even contain lighting that simulates a digital readout!

### the digital clock

Returning to my station and looking at the old reliable HT-37 and its companion receiver made me feel that something was lacking. However, the price tags on some of the newer gear made my equipment seem more acceptable. Still these electronic displays really caught my fancy — I *had* to have a piece of digital-display gear!

After some thought I came up with some requirements. My first digital display project had to be low in cost, simple, and small, in that order. My old 24-hour clock had given up the ghost after four years of faithful but loud service, so I decided on a digital clock for a first project. I'd seen a number of articles on digital clocks. These were not overly complex, but the component count was high, with many discrete components required in the display circuits. Most of these designs used a crystal time base operating at a high frequency and required a long countdown chain to divide the high-frequency signal to the frequency required by the clock circuit. To keep the component count down I looked into the use of the 60-Hz frequency as a time base. I found that, because of the power grid system used by the power companies to maintain the 60-Hz frequency, the clock accuracy over a long period would be nearly as good as that using a crystal time base. The short-term accuracy of the crystal system is better, but for a 24-hour clock for amateur use the 60-Hz line is a very good choice.

The next area for component reduction was the display system and the drivers usually associated with it. Most circuits use 11 driver transistors and as many as 22 resistors. After a little looking I found some display units that require only a simple four-line BCD input. The component count was coming down fast! You've all seen ads for the "Clock on a Chip," but a close look shows that it takes much more than just a chip. After some reading, shopping, and two evenings of wiring I got the clock down to two chips and no discrete transistors. Nothing fancy, but it functions well, is very easy to build, and looks great.

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The heart of the unit is the MM5312 IC (fig. 1). The internal functions of this chip are quite complex, but basically the chip divides the 60-Hz line frequency to one pulse per minute and advances its internal storage register at that rate. The output of that register is available in binary form at pins 1, 2, 3, and 24. The outputs are synchronized with the digit-enable outputs, pins 18, 19, 20 and 21. The coincidence of these signals allows you to apply the binary data to all four displays in parallel, with the digit-enable lines controlling which display accepts data at the correct time. This scheme minimizes the number of output lines required to operate the display. The second IC is a type SN7404N inverter. This chip converts the MM5312 binary output data to a TTL level for operation of the display units.

The display units actually contain ICs that store and decode the data and drive the LEDs that form the digital display. The displays have a blanking input, which we use to control their intensity in this application.

The power supply furnishes +5 and −12 volts to operate the circuitry and furnishes a 60-Hz reference for the clock chip through the 0.1 Meg resistor to pin 16 of U1. Zener-diode regulation maintains the output levels.

## construction

The digital clock was built using perf board and wire-wrap techniques since the circuit contains only two ICs. The perf-board and wire-wrap technique is fast and easy; best of all it is very forgiving. If you make an error or want to change a circuit, simply remove the wire and wrap on a new lead. The tool required for this technique is available from the source shown in the parts list, table 1. Number 30 AWG (0.25mm) wire is used with the wire-wrap tool, Strip insulation back about one inch (25mm) insert the wire into the tool, and place the tool over the pin to be wrapped. Turn the tool clockwise until the wrap is complete. This method provides excellent connections of the type used in many large-scale computer systems.

Layout and wire routing is not critical and any packaging scheme you might desire could be used. I cut a piece of perf-board slightly smaller than the inside dimensions of the case, then I cut a second piece of board large enough to hold the single socket used for the four displays. The smaller display board was then epoxied to the front edge of the larger board to form a right angle for the display, as shown in fig. 4. The

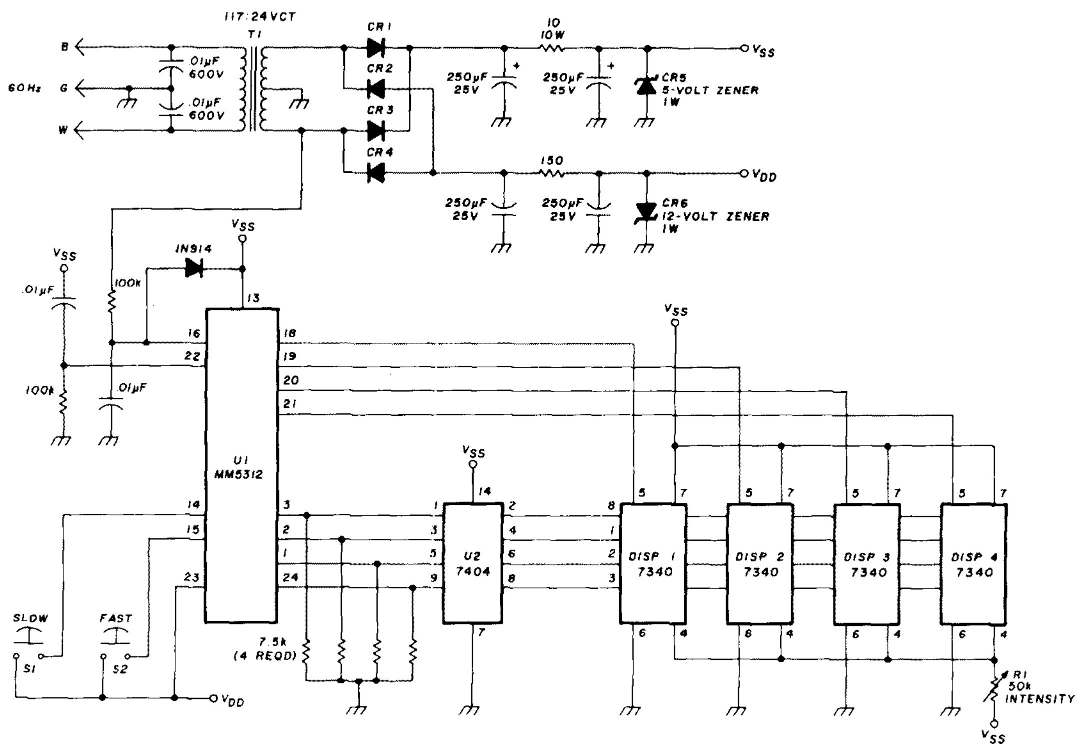
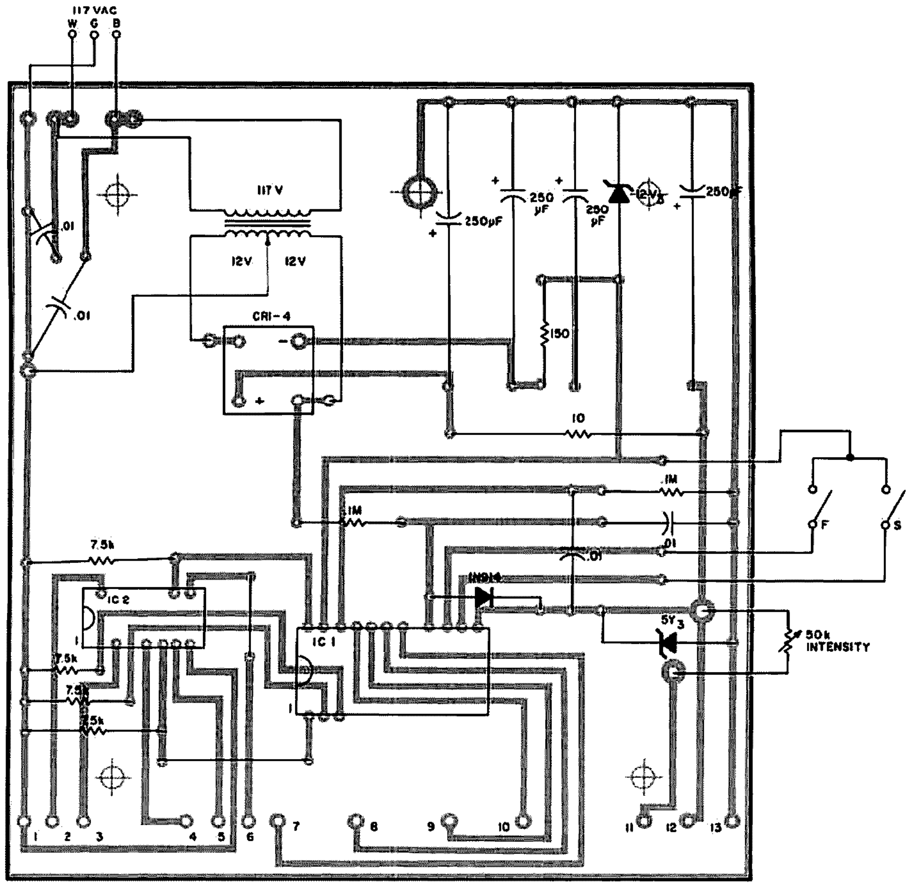



fig. 1. Schematic of the 24-hour digital-display clock including power supply. Resistors are ½ watt and capacitors are 50 volts or greater unless indicated otherwise.



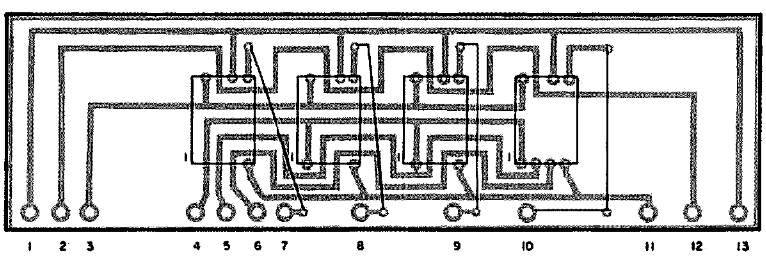




**fig. 2. Top view of the circuit board showing parts placement, and external wiring.**

display and IC sockets were mounted next. Fig. 5 shows pin layout and other useful information for the ICs and display devices.

The discrete component leads were bent at right angles and inserted into the board near their respective connection points. A drop of household cement will hold the discrete components to the board and make wiring easy. The wire wrap can be used on small-diameter component leads as well as on the socket pins; however, on the round leads of the resistors and capacitors it will pay you to put a little solder on the connection after wrapping to prevent intermittents later on. The larger leads, which are too large in diameter for wrapping, can be hand wired with the no. 30 AWG (0.25mm) wire, then soldered. After mounting all the components on the board the wiring can be completed in accordance with fig. 1. The leads for the fast and slow set switches and the dimmer control can be dressed out of the board and connected to their respective components after the board is mounted in the case. The board

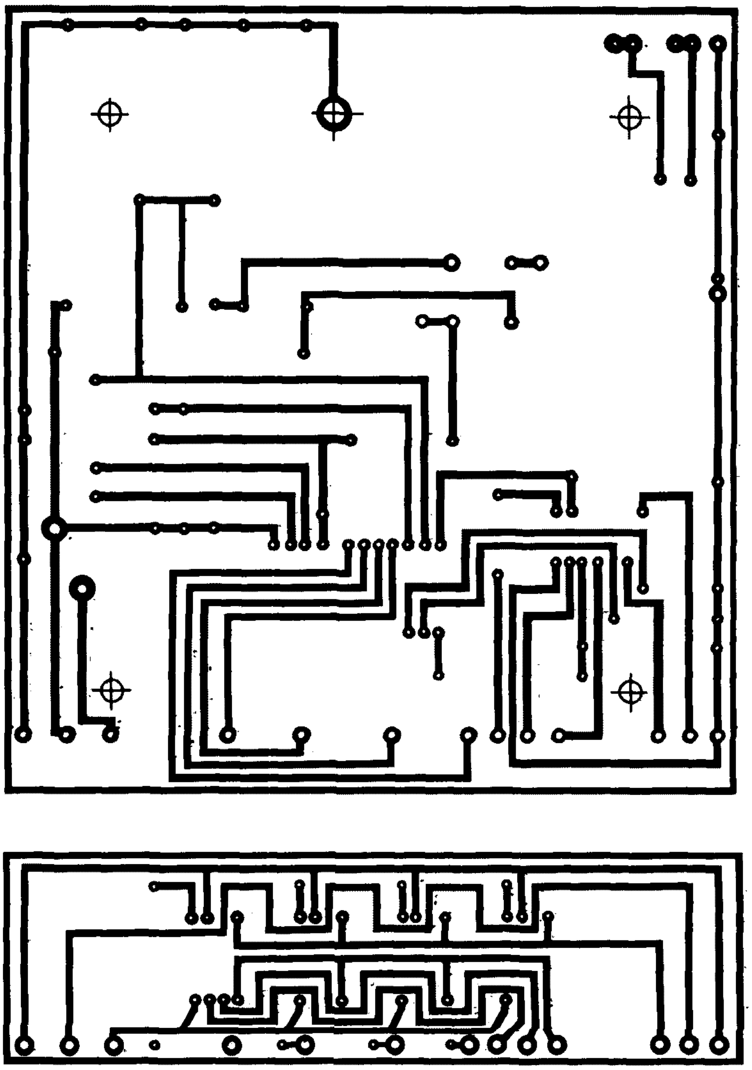


fig. 3. View of the circuit board from the foil side.

**table 1. Component list for the digital clock using perf-board and wire-wrap construction. Numbers in parentheses refer to product sources.**

| | |
|---|---|
| U1 | National MM5312N (3,5) |
| U2 | Texas Inst. SN7404N (2,5) |
| Display 1-4 | Hewlett Packard 5082-7340 (4) |
| T1 | SSS 11-08180 (1) or 273-1480 (2) |
| S1, S2 | 275-1547 (2) |
| R1 | 271-1716 (2) |
| CR1-4 | 276-1146 (2) |
| CR5 | 276-561 (2) |
| CR6 | 276-563 (2) |
| Socket U1 | SSS 41-38964 (1) |
| Socket U2 | SSS 41-38974 (1) |
| Socket, Display | SSS 41-38934 |
| Case | 270-253 (2) |
| Perf board | SSS 22-44062 (1) or 276-1583 (2) |
| Wrap Tool | SSS 43-18160 (1) |
| Bezel | 915-60 (6) |

**product sources**

(1) Solid State Systems, Box 773, Columbia, Missouri 65201.

(2) Radio Shack Stores.

(3) Poly-Paks, Box 942, Lynfield, Massachusetts 01940.

(4) Tinkers Electronics, 805D Union Ave., Los Gatos, California 95100.

(5) International Electronics Unlimited, Box 1708, Monterey, California 93940.

(6) Tracy Design, 15870 Schaefer, Detroit, Michigan 48227.

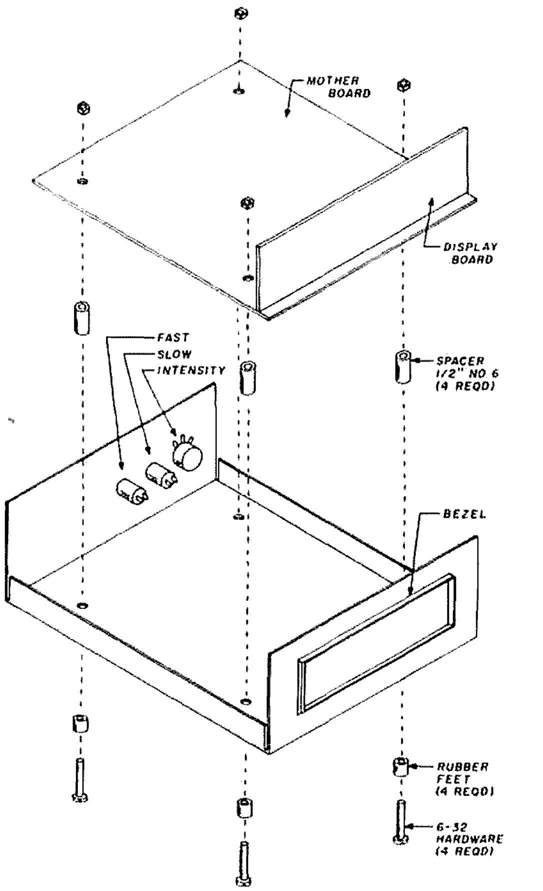
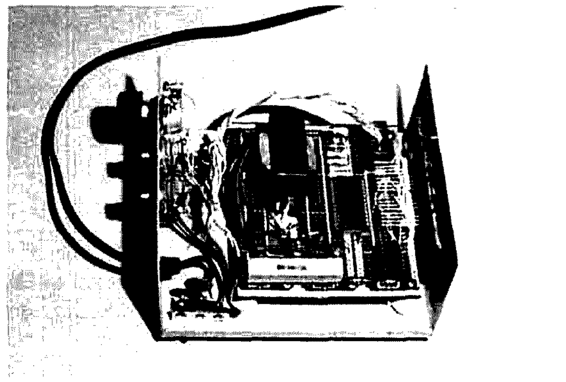



**fig. 4. Isometric of the 24-hour clock showing assembly details.**



**Top view of tne 24-hour digital-display clock with case removed.**

was supported from the bottom of the case with four spacers.

## checkout

After completing the wiring and before installing the ICs and displays, check the wiring against the schematic. The 0.100-inch (2.5mm) centers on the IC pins are very close and can be counted incorrectly quite easily. If all checks out well, plug in the ICs and the display units, apply power and give the circuit a smoke test. The initial display will probably scare you as some very odd configurations can appear before setting the correct time. The time is set by pressing the fast and slow set switches. The fast-set switch advances the clock at the rate of one hour per second. The slow-set switch advances the clock at the rate of one minute per second. Once set, you should have a very accurate 24-hour clock.

## conclusion

This project is a good starter in the transition from working with 807s to new gadgets like ICs and digital electronic displays, since the circuit requires no adjustment or elaborate test equipment. Parts are available at modest cost with a little shopping around, and the finished unit is a functional adjunct to your station. In shopping around for the display units I found that the reverse sides of the units had been painted black to obscure the HP part numbers, but holding the units up to the light made the part numbers easily readable. I found two other types of HP displays that also work in this unit, types 5082-7300 and 5082-7302. If these are used, pin 4 on each display must be grounded and the dimmer control eliminated. Devices that are functionally identical to the HP units can also be found under other manufacturer's part numbers.

If you're interested in the technical aspects of the ICs used in the clock, you should request data sheets when you purchase the ICs. The data sheets provide a very thorough explanation of the internal chip functions as well as very good application notes.

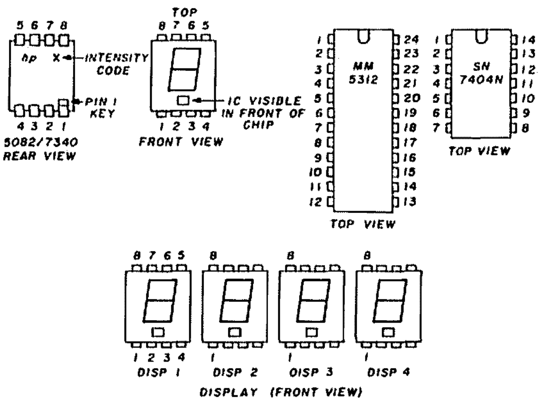



**fig. 5. Pin numbers of the ICs and display LEDs used in the 24-hour clock.**

This project has been a lot of fun and very educational. It has really stirred up my interest in digital techniques. What's on the work bench now? A frequency counter built in the same-size case as the clock. "Old Reliable" may soon have a digital dial.

**ham radio**

# audible S-meter for repeaters

## More audio tones on the repeater — this time they indicate signal strength

The **audible S-meter** can be used to check transmitter performance, antenna comparisons, site testing and so on, whenever relative signal levels are required. This unit, when connected to a repeater, will add a tone-burst shortly after the input signal has dropped out. During the "tail," the pitch of the tone indicates the strength of the preceding received signal. The S-meter can be built entirely from junk-box type components. In addition to the normal results, it adds a new experimental facility to any repeater.

### the repeater tail

Most fm repeaters have a "tail" period — the over run or run on time of the transmitter after the input signal has dropped out. It is assumed that the tail period is noise-free, *i.e.*, an unmodulated carrier. This time can be used to accomodate a toneburst or "beep" to telemeter the repeater receiver S-meter reading back to the original station.

The beep generated by this audible S-meter unit is about 60 milliseconds long — about the length of a time-signal "pip." It is long enough for the pitch of the signal to be understood and used, though not long enough to be annoying. It is possible to adjust the amplitude of the beep from zero to full deviation. The time delay between the input signal going off and the tone can also be adjusted from zero to about seven seconds, again to suit the requirements of the particular repeater.

The pitch of the resulting beep varies with the strength of the incoming signal. A high-pitched beep (3500 Hz) represents a weak signal, a low-pitched beep (350 Hz) represents a strong input signal. In between signals are reported accordingly. The pitch of the beep therefore varies inversely to signal strength. The high-pitch beep, for weak signals, is more useful to a distant station. If a low-pitch signal had been chosen for this condition it could be lost in the noise. Signal-to-noise considerations influenced the choice of this method of presentation.

This 10-to-1 change in pitch enables users to obtain the information without the necessity of installing accessory equipment to decode the beep. The human ear is quite good at sensing relative tones and the use of this unit is learned very quickly. Once installed, talk of its removal is soon squashed!

The audible S-meter has been installed at the Mount Climie Channel C repeater in New Zealand for some time and has been found to be very useful. Each beep was purposely timed to be 3 seconds into the tail and serves the additional purpose of setting a 3-second gap between stations to permit emergency or priority traffic time to break in between ordinary ham conversations. This 3-second delay is a local operating custom. The design can be modified to suit the requirements of other repeater groups and other regulatory bodies. It is described here in the hope that it may give inspiration to other experimenters.

### the repeater receiver

An S-meter must already be available at the repeater receiver. Just what form this takes will vary from repeater to repeater. This unit has been built around a receiver using an RCA CA3089E in its tail end — but the idea could be adapted to other receivers. The CA3089E has an S-meter output (called a "tuning indicator" on the data sheet) at pin 13. A 33k resistor and a 300 $\mu$A meter are normally connected to act as a tuning indicator. A switch changes the output over to the audible encoder as shown in the schematic, fig. 1.

Transistor Q1 is a dc amplifier, with its operating conditions and gain adjusted by R1 and R2. Q2 is an

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emitter follower, that provides a low-impedance charging circuit from its emitter to fast charge C1 (the memory capacitor) via diode CR1. A memory is necessary to store the information and hold it ready for broadcast after the input signal has gone off. Diode CR1 is

A tone-burst has to be generated from this continuous tone. Q8 acts as a clamp across C6 and prevents the audio oscillator from running constantly. The base of Q8 is forward biased and, as a result, Q8 is normally conducting. This transistor has to be shut off to produce the

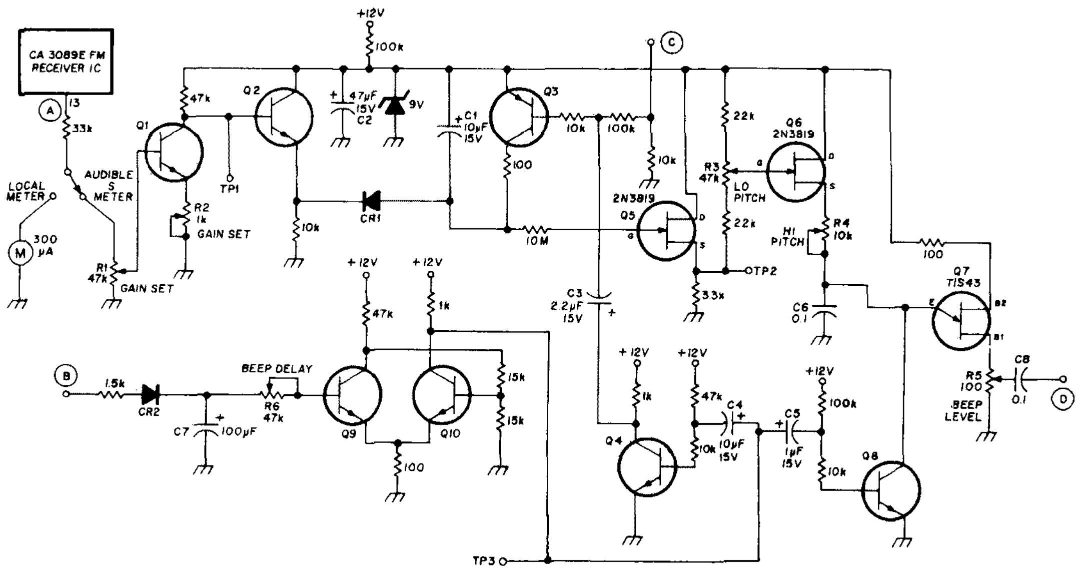


**fig. 1. Schematic diagram of the audible S-meter. Input B is connected to the squelch circuit. Point C is connected to a source of +12 volts that indicates when the transmitter is on. The tone output is available at point D. Changing the value of C5 will alter the length of the time "beep." Unless otherwise indicated the transistor types are discussed in the text.**

required to prevent the memory capacitor from discharging after the input signal has gone off.

## generating the tone

The voltage across the memory capacitor is "read" by the high impedance electronic voltmeter, Q5, acting as a source follower. Q5 effectively repeats the voltage on the capacitor but at a much lower impedance which can readily be metered for test purposes. The variable resistance of Q6 plus R4 act as the resistor portion of an RC network with C6 as the capacitor. Q7 is a unijunction transistor connected to act as an audio oscillator. The resulting tone from across R5 is applied to the repeater transmitter audio stages. As the effective resistance of Q6 changes, so does the frequency of the tone.

The operation of the unit as an S-meter current-to-audio tone converter can now be seen. Various levels of input current create a voltage across C1 and in turn cause corresponding tones from the audio oscillator. The high and low pitch ends of the audible operating range are set by R3 and R4. The output tone amplitude is controlled by R5. Because there is a standing dc current from pin 13 of the CA3089E, R1 is used to set the operating point of Q1 over the required current range only. R2 acts as a gain adjustment to set the required sensitivity of the system.

beep. This unclamping is achieved by Q9 and Q10 which form a Schmitt trigger circuit.

Lead B is connected to another part of the repeater receiver – the squelch and control system. This input is about +10 volts when a signal is being received and goes to ground immediately after the signal goes off. While the input is being received, C7 charges via CR2 to some positive value. Q9 is therefore conducting and Q10 is held cut-off so TP3 is at a high positive value.

When the received input signal goes off, point B goes to ground and C7 discharges via the base of Q9, maintaining Q9 on for the delay period. When C7 has discharged, as determined by R6, Q9 can no longer hold Q10 cut off and TP3 drops very suddenly to a low positive value, when Q10 starts conducting. This sudden drop is transferred through C5 to cutoff Q8. The tone oscillator can now operate until C5 has discharged. This takes about 60 ms – the beep length. We have now converted our tone into a beep which is of a desired length (selected by C5) and at an adjustable duration during the "tail" (set by R6).

## resetting

One problem remains. Some means must be provided to discharge the memory capacitor after it has been read, so it can recharge each time from the same initial voltage. Two discharge circuits perform this action, with Q3

acting as another clamp. When no receiver input signal is present, it places a short circuit across C1. As the input signal appears, the repeater transmitter turns on and point C goes to 12 volts, thus unclamping Q3. At the end of the tail period, Q3 conducts and discharges C1.

In addition, the negative-going voltage transition at TP3 is applied to Q4, cutting it off. Since Q3 is already unclamped because of the 12 volts on C, the positive pulse that comes through C3 only adds to keeping Q3 unclamped. When C4 has discharged, Q4 will come back into conduction. The negative-going pulse from the collector is transferred through C3 and clamps Q3, thus discharging the memory capacitor.

The two cancellation circuits are necessary to ensure the initial charging conditions of C1 are the same over long periods of time or over periods of little repeater activity, and also to ensure that capacitor C1 is discharged after every beep to allow for those cases where the repeater tail does not drop out before the next station begins transmission. We now have an audible beep which follows and stores the input current value on lead A and automatically resets after each readout.

## construction

Details pertaining to the construction of this unit are not considered to be necessary. It is one of those devices that is unlikely to be attempted by persons not already experienced in the noble art of wielding a soldering iron. The npn and pnp transistor types have purposely not been quoted. There is such a range of suitable substitutes that any one of many would be acceptable. The ones actually used were BC107 and BC177 of European origin. Any small-signal npn and pnp silicon transistors of medium gain should be suitable. Small signal silicon diodes should be used for low leakage.

## general

The unit in effect reads the *maximum* input signal received over a period because it is, in essence, a peak-reading voltmeter. Such an S-meter, relying on a 60 ms beep to send back information, can only operate on a sampling basis anyway. The maximum received strength is perhaps the most useful value to sample. Over a fading path this aspect must be remembered if useful interpretations are to be made. In practice, few problems in this regard occur because path conditions (except for mobile use) are surprisingly constant with fading occuring on a regular basis with each test.

A remote readout unit, that can be connected to transceivers, has been built. This puts the S-meter at an unusual end of the radio link — back at the transmitter where it is probably of much more usel A means of switching the encoder to another receiver at the repeater site for listening to a distant beacon station for propagation studies is also in operation. This too may bear telling at a future time. Meantime, this basic unit lets repeater users learn of their relative signal levels at the repeater receiver. I would like to know of any repeater installations where this or similar units are in use or brought into use as a result of this article.

**ham radio**

# repair bench

# Joe Carr, K4IPV

## basic troubleshooting: is that oscillator running?

Whether the discussion is among amateur or professional troubleshooters, one often-heard question is, "How do you tell if the oscillator is running?" In some cases this question is easy to answer, but in others you must resort to some reasonably clever troubleshooting. This is especially true when some of the possible methods that might be applied could easily cause oscillations to cease, in which case "no" is a hardly useful answer in all instances!

First, let's discuss some aspects of oscillator circuits. All such circuits have two subcircuits as part of their anatomy. These could be called the dc and the ac circuits. In the dc circuit, dc currents flow. The dc circuit includes elements such as bias resistors and emitter or cathode resistors. The ac path, on the other hand, contains feedback elements, crystals, tank circuits, and most (perhaps all) capacitors. These two paths are not mutually exclusive, as ac and dc components exist at many points. At the plate of a vacuum-tube amplifier, for example, a high dc voltage and an ac signal will be found. Similarly, if a stage is shunt fed the rf tank will be common to both ac and dc paths.

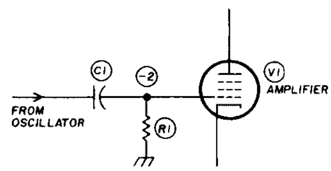


**fig. 1. Input circuit found in many small amateur transmitter amplifiers. Negative voltage on the grid of the amplifier tube provides a good indicator of oscillator operation.**

The point is that we must be careful when troubleshooting not to create a situation that gives a false result. Some test procedures, for example, may interfere with the operation or may only be valid for one path or the other.

Certain properties of oscillator circuits exist that cause dc voltages to occur only when the oscillator is running. Consider, for example, a simple grid-leak bias situation as in **fig. 1**. This is a common circuit of many transmitters of the master-oscillator, power-amplifier type. Assuming oscillator input voltage, the negative voltage on the V1 grid is caused by grid-leak action of R1, C1. Absence of this negative voltage can mean one of three thines: R1 open, C1 open, or the oscillator signal is missing. R1 can be checked by measuring its resistance with an ohmmeter. C1 can be also checked using an ohmmeter by following this simple procedure:

1. Disconnect one end from the circuit.

2. Short both capacitor leads briefly; then remove the short.

3. Connect the ohmmeter (-) to one end of the circuit.

4. Touch the ohmmeter probe to the other end of the capacitor.

If the capacitor is good, a brief downscale-flicker of the meter pointer will occur at the instant the probe touches the capacitor free end. The pointer will increase to "infinity" as the capacitor charges. With capacitors less than 0.001 µF, this flicker will be brief and shallow, so you may miss it during the first attempt. This test must be made using the highest-possible ohmmeter scale (X1 megohm is desirable). If the capacitor is leaky the ohmmeter will not return to infinity after the capacitor is charged but will hang at some value. In most circuits, this would have made the voltage at the grid of the amplifier tube positive instead of negative.

### checking variable-frequency oscillators

Many vfo and crystal oscillators will also present a

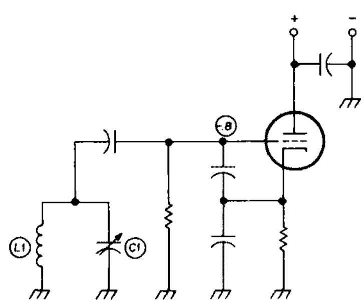


**fig. 2. LC-tuned fixed-frequency oscillator. L1, C1 can be short-circuited to check oscillator operation (not recommended for transistor circuits — see text).**

negative voltage due to grid-leak action at their *own* grids. Furthermore, this voltage exists only when the oscillator is running and will vary as the vfo is tuned through its range; what better indicator could you want? Of course, we must expect this variation to be roughly


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proportional to frequency change ($\Delta F$). In the vfo of an ssb transceiver, which covers only 250 kHz per bandswitch position, we'll find less variation in the dc grid potential than on a general-coverage receiver vfo, which covers a range of about 3.2:1. The variation will, however, be great enough to see on a sensitive dc voltmeter. This indication will tell you whether or not the local oscillator is running. If the stage is not oscillating, the grid voltage may drop to zero or will become a few millivolts positive.

## checking fixed-frequency oscillators

The oscillator grid-bias test is also useful with crystal or LC-tuned fixed-frequency oscillators. In these circuits it's necessary to disable the crystal (or LC tank) and look for an abrupt change in the bias level. Almost anything that will disable the oscillator or cause it to shift frequency by a large amount will do. You might want to try the following:

**1.** Touch the tank or crystal with a metal screwdriver. (Make *certain* that no high dc potentials exist at this point).

**2.** Remove power and remove or disconnect the crystal or one of the tank-circuit components.

**3.** Shunt a large-value capacitor across the crystal or tank. In most oscillators, anything in the range 0.001-0.05 $\mu$F will do.

**4.** Place a jumper wire across the crystal or tank circuit. Be certain of the circuit; *know* what current is being shorted and to where it is going. In **fig. 2** no damage will result if the tank is shorted to ground; but in at least one version of the Pierce circuit, and in some vfo circuits, this maneuver will cause further damage, especially in transistor circuits. A rule of thumb is: Avoid this method if B+, cathode, emitter, or base voltages share a common terminal with the crystal or tank.

Of these four methods, I prefer either **2** or **3** as those least fraught with danger. If any of these methods produces a sudden shift in or loss of the negative bias

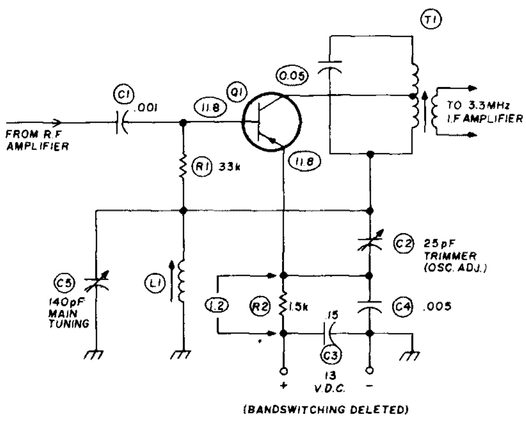



**fig. 3. General-coverage receiver converter vfo circuit. Such circuits can be checked by observing changes in emitter-to-base voltage.**

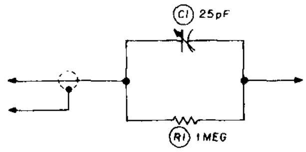



**fig. 5. Low-capacitance probe for use with an oscilloscope or digital frequency counter.**

voltage at the grid, then you can assume the osillator is running.

## checking converters

Fig. 3 shows a converter (oscillator and mixer combined in a single stage) as used in a general-coverage amateur receiver. Both emitter and base will be at the same dc potential when the oscillator is running: A good test point. If problems in the ac path cause the oscillator to stop running, then emitter and base voltage will

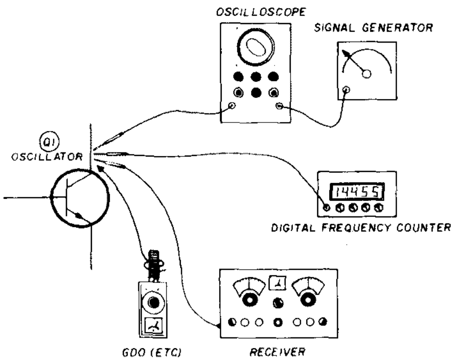



**fig. 4. Methods for checking oscillator frequency using common test equipment.**

change slightly, and the stage will act more like an amplifier. In this case, the base-emitter voltage will increase from zero to between 0.1 and 0.7 Vdc. Unfortunately, this test isn't always reliable but offers a good reason to become extremely suspicious. A better indicator of operation is the voltage drop across the emitter resistor, R2. This voltage always varies as the oscillator is tuned across its operating range. This method can also be used with fixed-tuned oscillators by disabling the crystal or tank and using the methods described above.

Before proceeding further, some cautionary words are in order against arbitrarily disabling an oscillator, especially in transmitters or other relatively high-power circuits where grid-leak bias is used: in many low- or medium-power transmitters, the grid leak provides the only bias for the final amplifier tube or tubes. If this bias is removed, the tube may be permanently damaged. In such cases it will be necessary to a) clamp the grid with a negative voltage high enough to cause cutoff, b) remove the tube or tubes, or c) disconnect the B+. Of these methods, the last is the most desirable.

Dc testing with a simple dc voltmeter does not always tell the tale, or at least the truth! In many cases, the dc

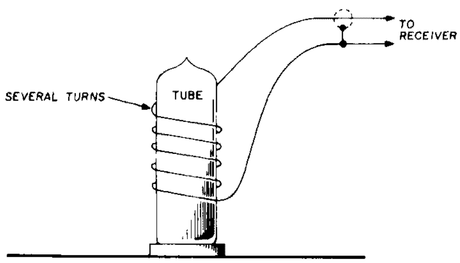



**fig. 6. Coupling method using the time-honored gimmick capacitor.**

indications merely cloud the issue and give a false sense of security. For example, how useful are the dc checks when the oscillator is still running but at an incorrect frequency? In that case, the dc checks will only lead you astray. Clearly, some method for measuring the oscillator frequency should be employed.

## checking oscillator frequency

There are several ways to rough-check the frequency of a running oscillator that usually do not cause any undesirable frequency shifts (**fig. 4**). One is to use a calibrated oscilloscope and measure the period of a single cycle of the oscillator signal by counting spaces on the scope time base (horizontal) and multiplying by the time base control setting. Or you could feed the oscillator signal to the vertical input and the output of a calibrated signal generator to the horizontal input. This action will create a Lissajous pattern. Adjust the signal generator until the trace remains steady and apply the normal rules for Lissajous figure ratios.

Alternatively, you could use a digital frequency counter or a heterodyne frequency meter. These instruments are available at low cost and many amateurs own them. It's also possible to find amateur clubs that have these instruments, especially those that operate repeaters.

Both the frequency-meter and Lissajous-pattern methods have a common fault, which may render them useless in some cases: Their probes may tend to load the circuit and shift its operating frequency. **Fig. 5** shows a low-capacitance probe. A grid-dip meter can be used as an absorption wavemeter, which can be loosely coupled to the circuit with a *gimmick* capacitor as shown in **figs. 4** and **6** or placed close to the oscillator.

Another method that has wide application is to use a general-coverage receiver. Connect the coax from the *gimmick* to the receiver antenna terminals and tune for the strong oscillator signal. In some cases it will be merely necessary to connect a length of hookup wire to the coax inner conductor and place it near the oscillator circuit.

The technique described above, which has a host of other uses, is one I recommend for every amateur station. Some of the old war horse receivers of the 50s and 60s are available for a price lower than you might think. Even an old ac-dc shortwave version of the five-tube superheterodyne is acceptable for many uses. An old Hallicrafters S38E or National NC-3 can be picked up at many swap meets for little money.

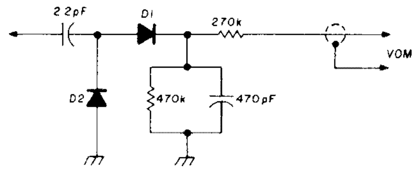
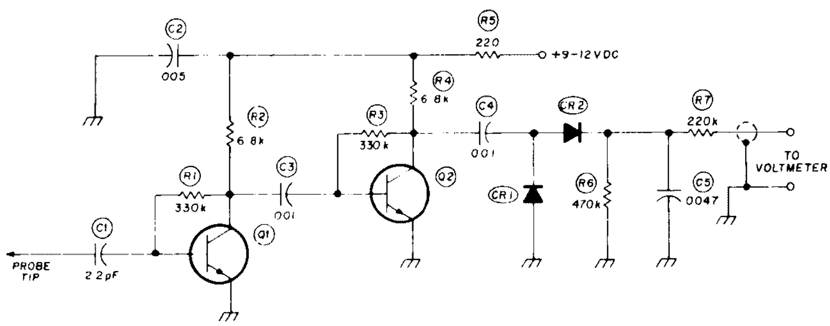


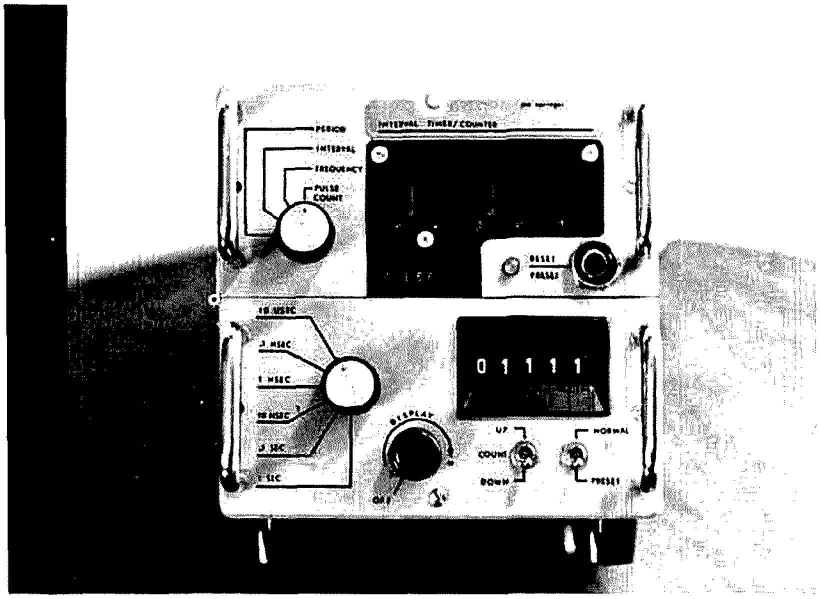

**fig. 7 Passive demodulator probe circuit, which can be built into an appropriate housing. CR1, CR2 are 1N60, 1N82, 1N914, etc.**

**Fig. 7** shows a passive probe for use with a vom. EICO sells probe kits as well as demodulator probes, or you can buy a blank kit and build your own probe into the barrel. This is the approach recommended for the active probe of **fig. 8**. This demodulator probe is useful to two meters or higher. Transistors Q1 and Q2 are an amplifier that feeds a voltage-doubler detector consisting of CR1, CR2. The probe should be built into the same housing as the probe tip to eliminate lead inductance between tip and capacitor C1. The EICO housings can be used, or an old probe can be used. One-half inch (12.5mm) copper tubing with end caps is also useful, as are 9-pin tube sockets with shields (see the ARRL *Radio Amateur's Handbook*).

**fig. 8. Active demodulator probe, which is useful to 2 meters or so. Q1, Q2 are hf npn transisitors (ECG108, etc.) CR1, CR2 are 1N82, 1N914, etc.**




**ham radio**




# function/units indicator

## using LED displays

Design and construction of a simple alphanumeric display system with many applications

Every test instrument uses some form of front-panel display to indicate what is being measured and the units involved. This article provides an attractive alternative to conventional displays. The design uses LEDs and TTL ICs to indicate function and units in an interval timer/counter. However, the circuit is adaptable to a wide range of other instruments.

The display complements the timer/counter and makes reading of measured data easier than with instruments using conventional display methods. Time in milliseconds, for example, is displayed as *SEC-3* (seconds x $10^{-3}$) in scientific notation. Frequency is displayed similarly. The indicator circuits use TTL ICs, the heart of the system being programmable read-only memory (PROM). Construction is straightforward. A simple programmer for the Signetics 8223 PROM has also been constructed and included as part of the project.

After a brief discussion of concepts and theory of operation, the design techniques I used are described as well as the evolution of the final circuit. Construction is covered in detail for those who may wish to duplicate the circuits.

### function/units indicators

The purpose of a function/units indicator on an interval timer/counter is to describe what is being measured: seconds, milliseconds, or microseconds for time measurement and hertz, kilohertz, or megahertz for frequency

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measurement. Most instruments have abbreviations of these units (sec, msec, μsec, Hz, kHz, MHz, respectively) on the front panel and use either a pointer or panel light to indicate the appropriate display.

The LED indicator displays any one of the six abbreviations with a minimum of five 7-segment digits placed side-by-side. Seven-segment displays are designed to display numbers only; fortunately, enough letters of the alphabet can be produced to simulate the abbreviations. (See table 1). Obviously, this method is more complex: it requires considerable circuitry and the cost is higher. Is it practical? Economically, no; but the greater ease in reading the instrument justifies its use.

Circuit operation centers around a 256-bit read-only memory, a Signetics type 8223,[1] into which are programmed all letters and numerals required to form the six abbreviations. One exception exists. Since there was no need to display microseconds (*SEC-6*) in my interval timer/counter, I used that section of memory to

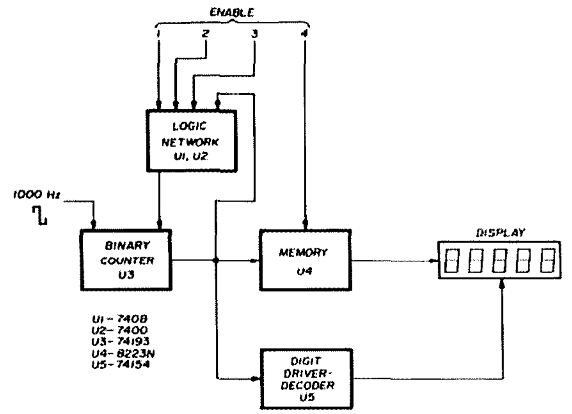
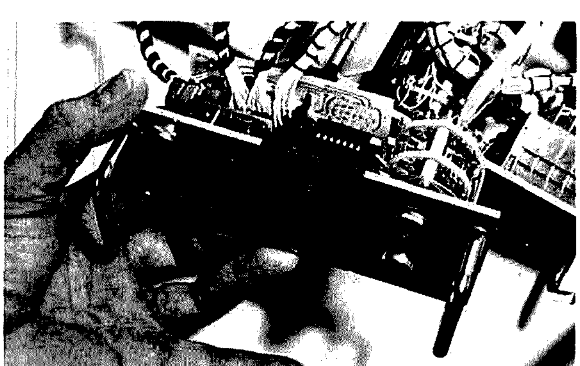



**fig. 1. Functional block diagram of the function/units indicator. A binary counter, a 4-16 line decoder, and a 5-digit parallel-connected display form a simple multiplexer. An external 1-kHz oscillator drives the counter and sets the scanning rate.**

**table 1. Alphanumeric simulation of frequency and time units.**

| units nomenclature | display |
|---|---|
| seconds | SEC |
| milliseconds | SEC-3 |
| microseconds | SEC-6 |
| hertz | HZ |
| kilohertz | HZ 3 |
| megahertz | HZ 6 |

program the word *PULSE*. This word is selected for display when the counter is set up to count random pulses or events. The truth table for the memory is provided in table 2. A block diagram of the indicator circuit major units is shown in fig. 1. A binary counter, a 4-16 line decoder, and a 5-digit parallel-connected display form a simple multiplexer that addresses memory, one word at a time, and transfers data to the display. An external 1000-Hz oscillator drives the counter and sets the scanning rate. The counter is controlled by a logic network, which in turn is controlled by three data lines.



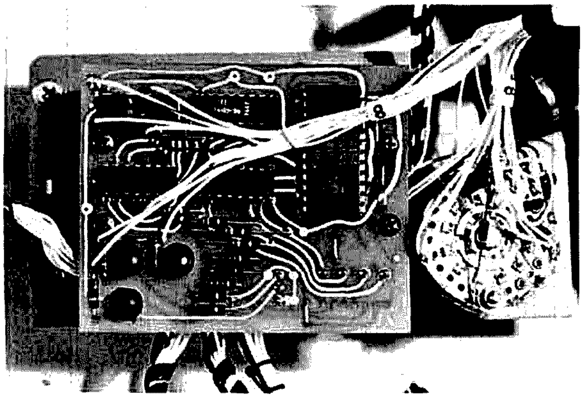
**Top view of the function/units indicator.**

The logic network limits the counter to count from 0-4 and reset; or 5-9 and reset; or 10-14 and reset – depending on the input from the data lines. Enable 4 is the most-significant bit address line to the PROM and selects one of two blocks of memory, locations 0-15 (binary) or 16-31 (binary). A schematic of the indicator circuit appears in fig. 2. Table 3 lists the data inputs for each different display.

## design

Four areas of special importance were stressed in the design: small size, simplicity of circuit (minimal inputs), parts availability (cost included), and symmetry (printed circuit layout). Small size (3.2 x 2.5 inches or 8 x 2.5cm) was necessary only for my application. This circuit would probably function quite well under almost any configuration. Lead lengths should be kept as short as possible, however, in the interest of good design practice. Simplicity is the keynote for any good design and merited a major effort in this project. All parts are readily available on the surplus market. Table 4 gives information on parts and suppliers. Cost is within reason ($12.00) if surplus ICs are used throughout. Special attention was given to PC layout, since two-sided foil patterns were necessary and feedback problems might be encountered.

**input and output requirements.** The function indicator is controlled and operated from the rotary function switch on the main instrument panel through the data-input lines. As table 3 indicates, each of the six displayed outputs requires a different 4-bit binary code. TTL integrated circuits are used exclusively, so 5 volts dc is the only power source necessary. A 1000-Hz square wave is provided by the main instrument's time-base generator to drive the multiplexer. The square-wave frequency is arbitrary; however, a minimum of approximately 300 Hz is necessary to prevent noticeable blinking of the dis-



**Bottom view showing components and wiring harness arrangement.**

play. Input signal levels must be compatible with standard TTL drive requirements.[2] The five-digit display is the only output. The display driver, U3, limits the maximum static display sink current to 16 mA per digit.

design techniques. Simplicity was foremost in arriving at a workable circuit. I concentrated on the logic network fig. 1, since it determined how many input-data lines would be required. The counter needed four signals from the logic network for proper sequence. (See fig. 2 for a complete schematic of the logic network). With some trial and error I constructed four circuits, one for each signal, then, borrowing on the techniques of tabular minimization and NAND-NOR implementation,[3] I combined the circuits into a single, minimal circuit.

TTL systems are inherently noisy[2] so some bypassing was necessary. A minimum of 0.1 $\mu$F mounted close to the U3 and U4 terminals, where most of the switching noise is generated, is adequate.

The binary counter and logic network form a feedback system, but since there are at least three levels of gate delay from input to output, no interaction was anticipated. The counter output drives only three other inputs, so the counter is well within its fan-out limitations. The values of the pull-up resistors on the memory-output lines were optimized at 680 ohms, ¼ watt, for best display brilliance.

programmable read-only-memory. Now that I had a workable circuit on the drawing board, I set it aside to tackle the job of programming the read-only memory. Commercial programming is available from several companies (table 4), but I elected to avoid the additional cost and built a programmer I could use in future projects. Signetics Corporation provides programming procedures in their data book.[1] I used this information to construct the circuit of fig. 3.

A word of caution is in order, however, about the use of *this programmer for* memories of different manufacture. The Texas Instruments type 74188A field-programmable memory[4] is an identical pin-for-pin replacement to the Signetics type 8223, but it *cannot* be

**table 2. Truth table for the Signetics 8223 ROM.**

| decimal | a4 | a3 | a2 | a1 | a0 | B0 | B1 | B2 | B3 | B4 | B5 | B6 | B7 | |
|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|
| 0 | 0 | 0 | 0 | 0 | 0 | 1 | 0 | 1 | 1 | 0 | 1 | 1 | 0 | S |
| 1 | 0 | 0 | 0 | 0 | 1 | 1 | 0 | 0 | 1 | 1 | 1 | 1 | 0 | E |
| 2 | 0 | 0 | 0 | 1 | 0 | 1 | 0 | 0 | 1 | 1 | 1 | 0 | 0 | C |
| 3 | 0 | 0 | 0 | 1 | 1 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | |
| 4 | 0 | 0 | 1 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | |
| 5 | 0 | 0 | 1 | 0 | 1 | 1 | 0 | 1 | 1 | 0 | 1 | 1 | 0 | S |
| 6 | 0 | 0 | 1 | 1 | 0 | 1 | 0 | 0 | 1 | 1 | 1 | 1 | 0 | E |
| 7 | 0 | 0 | 1 | 1 | 1 | 1 | 0 | 0 | 1 | 1 | 1 | 0 | 0 | C |
| 8 | 0 | 1 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 1 | 0 | — |
| 9 | 0 | 1 | 0 | 0 | 1 | 1 | 1 | 1 | 1 | 0 | 0 | 1 | 0 | 3 |
| 10 | 0 | 1 | 0 | 1 | 0 | 1 | 1 | 0 | 0 | 1 | 1 | 1 | 0 | P |
| 11 | 0 | 1 | 0 | 1 | 1 | 0 | 1 | 1 | 1 | 1 | 1 | 0 | 0 | U |
| 12 | 0 | 1 | 1 | 0 | 0 | 0 | 0 | 0 | 1 | 1 | 1 | 0 | 0 | L |
| 13 | 0 | 1 | 1 | 0 | 1 | 1 | 0 | 1 | 1 | 0 | 1 | 1 | 0 | S |
| 14 | 0 | 1 | 1 | 1 | 0 | 1 | 0 | 0 | 1 | 1 | 1 | 1 | 0 | E |
| 15 | 0 | 1 | 1 | 1 | 1 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | |
| 16 | 1 | 0 | 0 | 0 | 0 | 0 | 1 | 1 | 0 | 1 | 1 | 1 | 0 | H |
| 17 | 1 | 0 | 0 | 0 | 1 | 1 | 1 | 0 | 1 | 1 | 0 | 1 | 0 | Z |
| 18 | 1 | 0 | 0 | 1 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | |
| 19 | 1 | 0 | 0 | 1 | 1 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | |
| 20 | 1 | 0 | 1 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | |
| 21 | 1 | 0 | 1 | 0 | 1 | 0 | 1 | 1 | 0 | 1 | 1 | 1 | 0 | H |
| 22 | 1 | 0 | 1 | 1 | 0 | 1 | 1 | 0 | 1 | 1 | 0 | 1 | 0 | Z |
| 23 | 1 | 0 | 1 | 1 | 1 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | |
| 24 | 1 | 1 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | |
| 25 | 1 | 1 | 0 | 0 | 1 | 1 | 1 | 1 | 1 | 0 | 0 | 1 | 0 | 3 |
| 26 | 1 | 1 | 0 | 1 | 0 | 0 | 1 | 1 | 0 | 1 | 1 | 1 | 0 | H |
| 27 | 1 | 1 | 0 | 1 | 1 | 1 | 1 | 0 | 1 | 1 | 0 | 1 | 0 | Z |
| 28 | 1 | 1 | 1 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | |
| 29 | 1 | 1 | 1 | 0 | 1 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | |
| 30 | 1 | 1 | 1 | 1 | 0 | 0 | 0 | 1 | 1 | 1 | 1 | 1 | 0 | 6 |
| 31 | 1 | 1 | 1 | 1 | 1 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | |

Positive Logic: 1 = hi level 0 = lo level

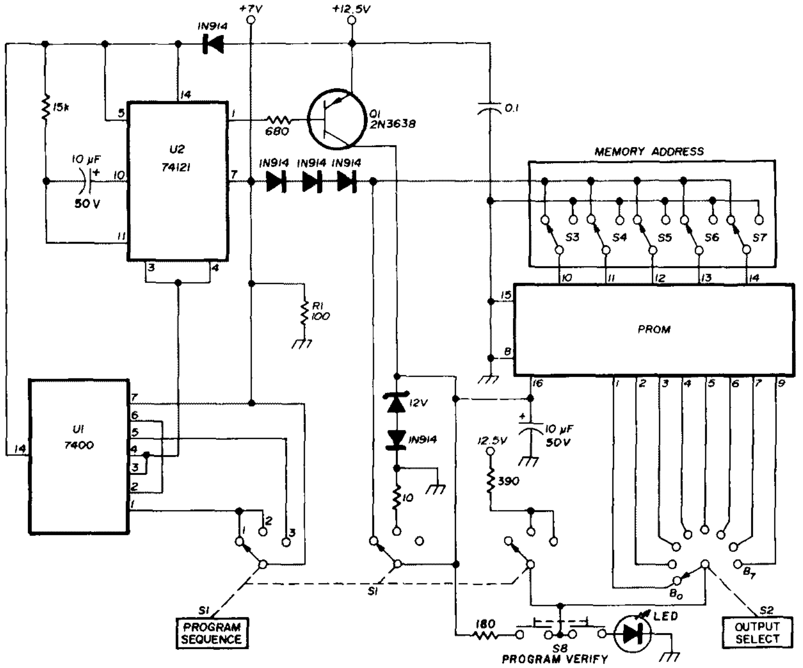



**fig. 3. Programmer using the Signetics 8223 IC. Circuit was designed and constructed from information in reference 1. The LED is an MV5753.**

programmed by the same procedures. Texas Instruments, Inc. has outlined special instructions.[4]

This programmer is much more elaborate than required but is designed to prevent accidental damage to the device being programmed. It also provides for checking the state of each bit of memory after each programming step. I'll not deal in depth on the theory of operation except for a brief overview. A complete discussion of the operating procedures is presented for those who may want to build this programmer.

Referring to fig. 3, U1 functions as a bounceless switch to trigger U2, a monostable multivibrator. Both

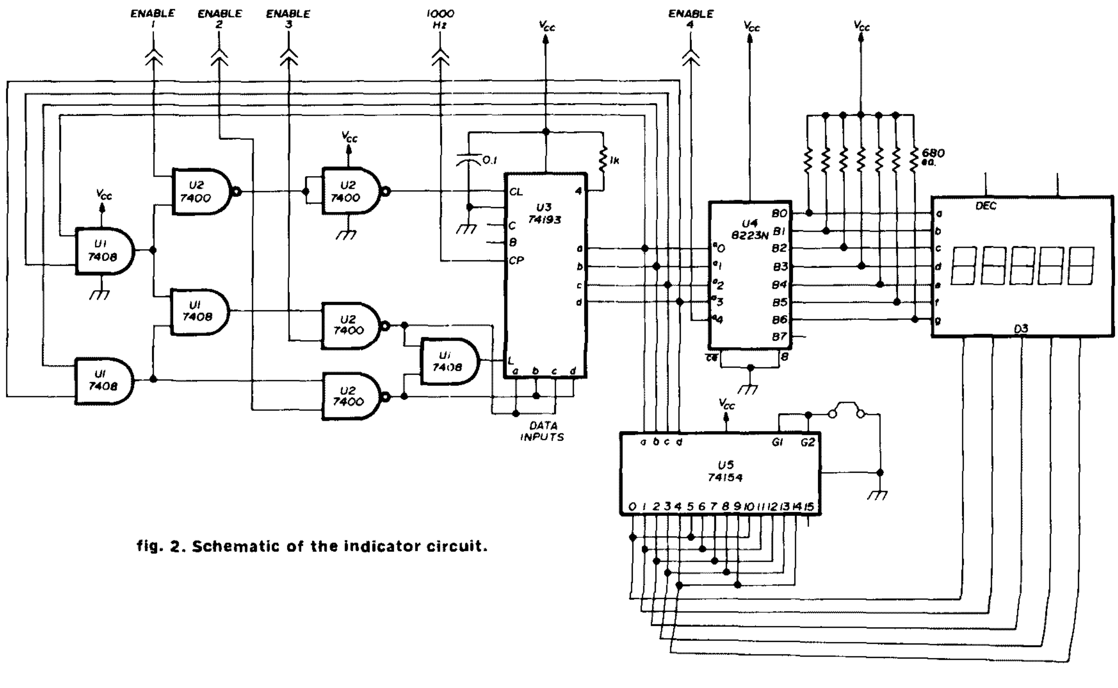



**fig. 2. Schematic of the indicator circuit.**

U1 and U2 operate at 7 volts above ground. Transistor Q1 saturates upon receiving a pulse from U2 and applies a 250-ms, 12.5-volt programming pulse to the $V_{cc}$ terminal of the memory chip. This pulse opens a fuse at the bit previously addressed and permanently establishes a logic 1 at that bit location. All memory locations are at logic zero before programming occurs. Separate regulated supplies provide the 7- and 12.5-volt inputs. Resistor R1 at the 7-volt terminal increases current and prevents interaction between the two power supplies.

**table 3. Data inputs (positive logic).**

| enable | | | | display |
|---|---|---|---|---|
| 1 | 2 | 3 | 4 | |
| 1 | 0 | 0 | 0 | SEC |
| 0 | 1 | 0 | 0 | SEC-3 |
| 0 | 0 | 1 | 0 | PULSE |
| 1 | 0 | 0 | 1 | HZ |
| 0 | 1 | 0 | 1 | HZ 3 |
| 0 | 0 | 1 | 1 | HZ b |

Programmer operation follows a simple, concise procedure that must be strictly adhered to. Read the following instructions several times and become familiar with them. A few dry runs are an excellent way to avoid a costly mistake. *The steps must be followed exactly as listed.*

1. Set switch S1 to position 1.

2. Address desired memory location (0-31, binary) with switches S3 through S7.

3. Select the output to be programmed (B0-B7) with switch S2.

4. Sequence switch S1 through positions 1, 2, and 3, and back to 1. (Complete the sequence in approximately two seconds).

5. Ensure switch S1 is in position 1. Then verify the bit programmed by pressing switch S8 momentarily. The LED will light if a logic 1 state exists.

   An erroneous reading will occur during verification if switch S1 is not in position 1.

6. Repeat steps 3, 4 and 5 for each output; (B0, B1, B2, B3, B4, B5, B6, B7).

7. This completes the programming of 1 of 32 memory addresses. Return to step 1 and continue.

## construction

The function/units indicator circuit is built on glass-epoxy base PC board, copper foil on both sides. A low-heat soldering pencil is absolutely necessary because of the density of circuit components and heat sensitivity of the ICs. A heat sink should be used for all soldering operations. In addition, I recommend tin plating the copper board before soldering components: it makes soldering that is otherwise tedious nearly effortless. It also adds a professional touch to the work.

Etching the circuit board is most efficiently done with the photographic technique, since accurate alignment of the two foil patterns is very critical. The two foil patterns, drawn to scale, are shown in **fig. 4**.

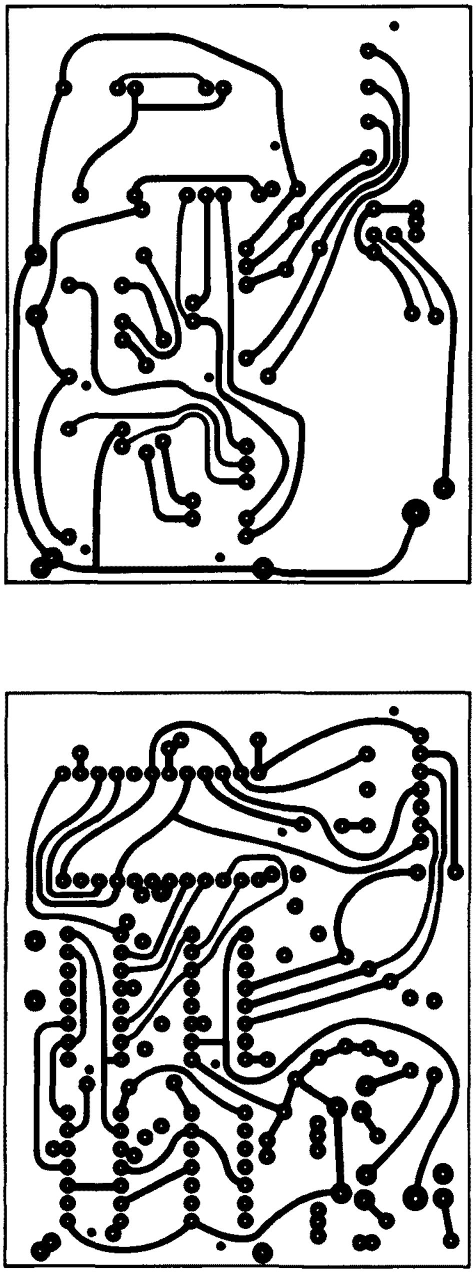


**fig. 4. Full-size PC-board patterns for the function/units indicator. Two-sided boards are used; photo-etching technique was employed to make the boards. The top board is the component side foil, while the bottom is the rear foil.**

Fig. 5 illustrates component mounting. Notice that the display is mounted on the opposite side of the board from the other components. This layout contains some other circuitry not related to this project, which can be ignored.

## performance

Preliminary testing of the project revealed two small problems, both with the data-input lines. Some noise spikes are apparent at the data (enable) inputs, and are generated by external switch contacts during switching. No effort was made, however, to remedy this situation since it was very minor. The other problem was a design error that caused a considerable waste of power. I'll explain. Proper circuit operation requires that the data-input lines, unless activated by an enable (high-level) signal, be at logic zero (low-level). To accomplish this, I had to clamp each data line to a low level with a 330-ohm resistor to ground. An unterminated gate, you recall, is at logic 1. A better design would clamp each line high with a 1000-ohm resistor and activate with a low-level enable signal. The current drain on the data lines would then be reduced from 60 to 20 mA. This is highly significant, since the entire function indicator draws only 220 mA at 5 volts.

The completed project was placed in a closed space and left on for two days in a rather makeshift test of its durability. No apparent heat problems occurred and the indicator functioned admirably. It is a handsome complement to my interval timer/counter.

The circuit is usable in a wide range of other instruments. The memory can be programmed for virtually any display configuration. I hope that this circuit, or portions of it, will stimulate some ideas for new applications.

**table 4. Parts and ordering information for the function/units indicator.**

**Integrated Circuits:** The following list contains just a few of the many dealers in surplus components

| | |
|---|---|
| Solid State Sales<br>P. O. Box 74A<br>Somerville, Massachusetts 02143 | Digi-Key Corporation<br>P. O. Box 126<br>Thief River Falls, Minnesota 56701 |
| International Electronics, Unltd.<br>P.O. Box 1708<br>Monterey, California 93940 | James Electronics<br>P. O. Box 822<br>Belmont, California 94002 |

**Displays:** Hewlett-Packard type, 5-digit, mounted on a 14-pin DIP socket

| | |
|---|---|
| Radio Shack<br>P. O. Box 2625<br>Fort Worth, Texas 76101<br>(see your local distributor) | Poly Paks<br>P. O. Box 942R<br>Lynnfield, Massachusetts 01940 |

**Programming Services:** (fees are generally around $5.00)

Semiconductor Specialists
Box 66125 O'Hare Airport
Chicago, Illinois 60666

**Available Programmers:**

Curtis Electro-Devices programmer
Spectrum Dynamics I/O programmer
Data I/O programmer

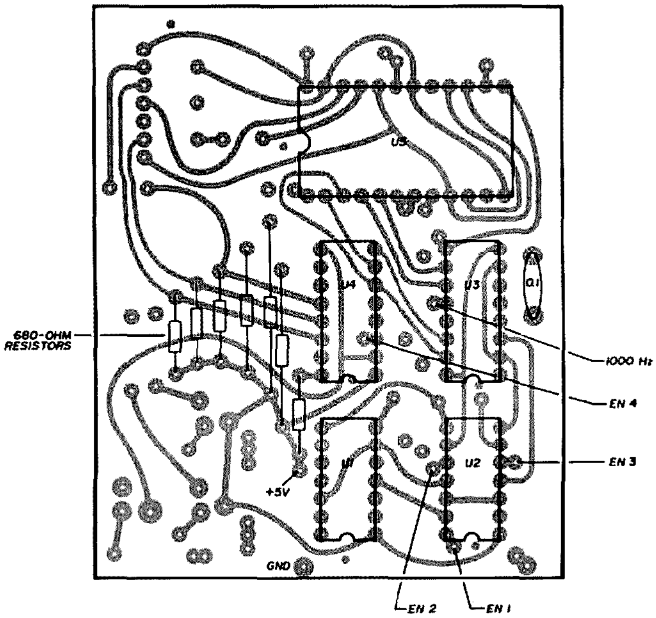



**fig. 5. Component layout for the indicator. Some of the circuitry is not related to this project and should be ignored.**

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3. W.J. Evanzia, "A New Wave of Test Instruments," *Popular Electronics*, 6, No. 4, October, 1974.
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**ham radio**

# control function decoder

## Repeater accessory access can be regulated by this control function decoder

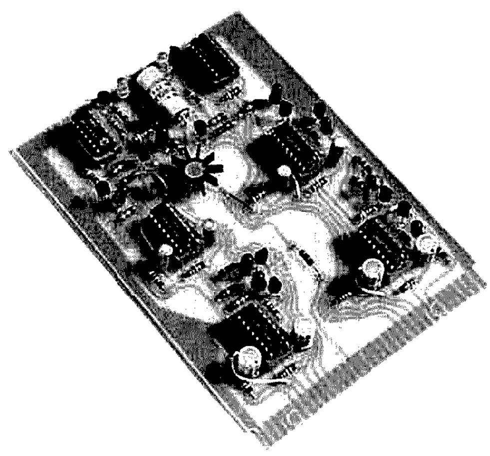
As repeaters become more sophisticated, the need for remote control of special repeater functions becomes necessary. Switching to remote receivers, autopatch control, and selectable repeater linking are only a few of the special repeater functions that lend themselves to remote control. The *Touch-Tone* signaling system is rapidly becoming the standard for repeater control. The control function decoder described here is designed to be operated from the output of a *Touch-Tone* decoder. The circuit detects a 3 digit sequence of *Touch-Tone* signals and sets a flip-flop. Another three-digit sequence is required to reset the flip-flop. The first two digits are the same for the set and reset functions. For example, the sequence 1, 2, 3 could be used to turn the function on, the sequence 1, 2, 4 could be used to turn the function off. The set and reset sequences must be received in the proper order and completed within a prescribed length of time.

The control function decode has several features that make it adaptable to a wide variety of systems.

1. On-card voltage regulation
2. Printed-circuit board construction
3. Plug-in module construction
4. CMOS chips for low power consumption
5. Fully TTL compatible inputs and outputs
6. Indicator LEDs monitor circuit functions

### circuit description

The circuit shown in fig. 1 uses two dual monostable multivibrators. The output of the first monostable (U1A) is connected to the reset terminal of the second (U1B). The first digit in the access sequence is the trigger



The complete control function decoder. The status of the timers is given by the light-emitting diodes mounted on the board.

By Thomas E. Doyle, WA9FTH, 5222 Big Bow Road, Madison, Wisconsin 53711

for the first monostable and the second digit of the access sequence is the trigger for the second monostable. Until the first monostable is triggered, the second is held in the mode. The output of the second monostable is connected to the reset terminals of the two remaining monostable multivibrators (U2A and U2B). The third digit in the *on* sequence is connected to the trigger input of one of the monostables (U2A) and the third digit in the *off* sequence is connected to the trigger input of the other (U2B). Note that the first two digits are the same for the *on* and *off* sequences.

The output of the *on* monostable is connected to the set input of an RS flip-flop. The output of the *off* monostable is connected to the reset input of the RS flip-flop. U1A is triggered by the first digit in the access sequence. U1B must be triggered by the second digit of the access sequence before the first monostable times out. The *on* or *off* digit of the access code must trigger its associated monostable before times out.

## interfacing

The output of the RS flip-flop is buffered by a transistor amplifier. An auxiliary direct clear input is provided to the RS flip-flop. This allows power-on clear or manual override. LED indicators are provided on all monostable outputs and the RS flip-flop output. The access sequence digits and clear inputs require positive going TTL level signals. The output voltage from the RS flip-flop is approximately 4 volts at a current level limited by R3. This resistor value may be adjusted dependent on output loading. The circuitry is operated from a 5-volt supply regulated by an LM309H regulator. This allows the decoder to interface with TTL circuitry. The power input should be in the range of 11 to 15 volts dc.

The R1, C1, and R2, C2 values control the length of time in which the three-digit sequence must be completed. The value of these components may be changed if desired. One of these boards is used in the WR9ABT 16-76 repeater to control the autopatch and also the repeater mode from carrier to tone access.

## construction

The control function decoder is built on a 4.5 x 6-inch (11.4x15.2cm) single-sided printed-circuit board. The board contains all the necessary circuitry for two complete decoders. These sections are completely separate and share only the voltage regulator. If only one decoder is required, the components for the other can be deleted.

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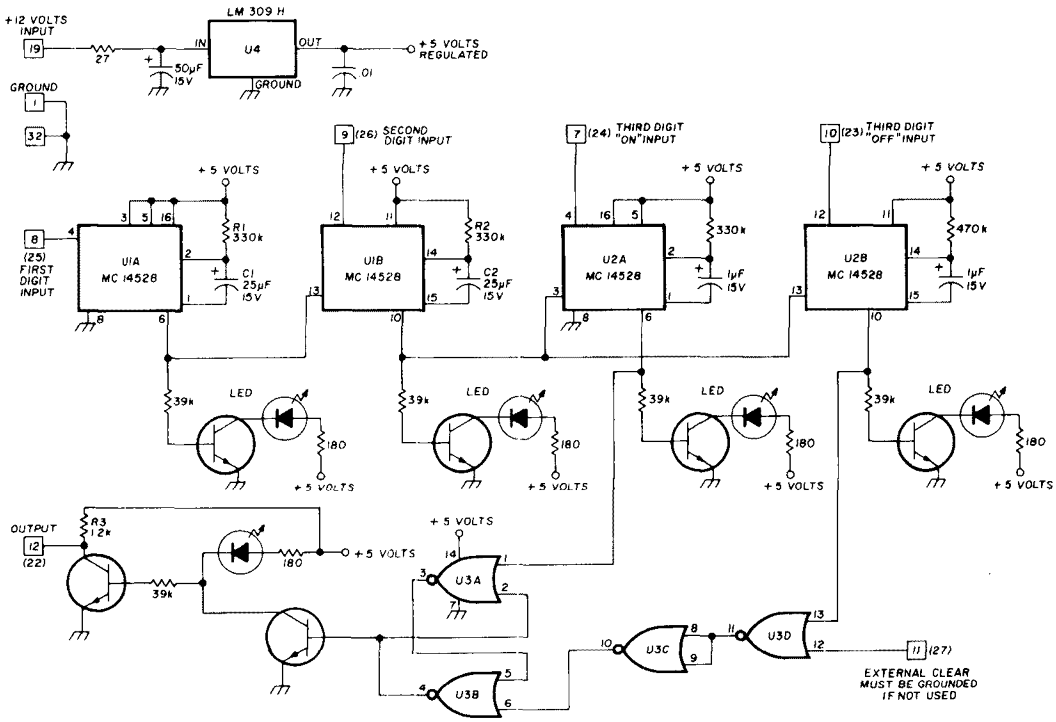
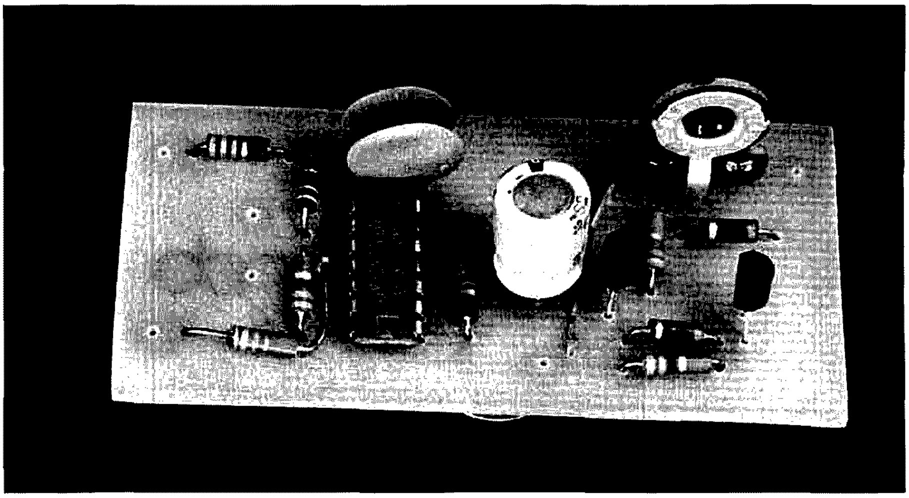



**fig. 1. Schematic diagram of the control function decoder. Q1 through Q6 can be most any npn silicon transistor (2N3904). The LM309H and associated heatsink are mounted on the circuit board. U3 can be an MC14001 or CD4001. The numbers in the boxes and parentheses refer to the edge connector pins. Note that there are two complete decoders on a single printed-circuit board. All resistors are ¼ watt, 10%. A copy of the printed-circuit layout is available from HAM RADIO upon receipt of an sase.**



# synthesized channel scanning

## Simple and easy channel scanning for the GLB 400 synthesizer

After acquiring one of the GLB 400B Channelizers, I found that it would be convenient to monitor our local repeater while hunting for others on the band. WAØAQO's[1] system was far too complicated for me so I incorporated a simple flip-flop to replace the function of the receive switch and presto, I had a two-channel scanner that would alternate between the two frequencies set into the Channelizer by the switches.

### circuit description

As shown in fig. 1, gates 1 and 2 are connected as an astable multivibrator with its frequency controlled by R6 and C1. Lowering the value of C1 will cause the oscillator frequency and consequently, the scan frequency to increase. Conversely, the scan frequency will be slower if the value of C1 is decreased. The square wave produced by this oscillator is used to drive gates 3 and 4, connected as a dual flip-flop.

The basic action of a flip-flop has one output high while the other is low. With each incoming pulse from the oscillator, the outputs will change to the opposite state. The high output of gates 3 and 4 is used to alternately select the frequency determining switches.

Now that the channelizer is scanning the two frequencies indicated by the settings of your switches, a method is needed to stop the oscillator when a signal is received. Note that fig. 1 shows two circuits for stopping the oscillator. Check your receiver to determine whether the point you desire to tie onto for scan lock goes positive or negative. A good place to tie in is the area of the squelch or limiter circuit. If this point goes in a negative direction when a signal is received, then use the circuit A. If the point goes positive, use circuit B.

In lock circuit A, Q1 is normally conducting and will cease conducting when a negative voltage or ground is applied. A positive voltage will then forward bias the base of Q2, effectively shorting gate 1 to ground and stopping the oscillator. Gates 3 and 4 will then remain in their respective states.

Q2 in lock circuit B operates in the same manner as circuit A. It is initially cut off until there is an incoming signal. Switch S1 is used to turn the Channel Scanner off and return the Channelizer to its original configuration.

### construction

The circuit can be built on a piece of perforated board* and mounted anywhere within your Channelizer that space permits. Three additional holes are needed in the front panel. Locate a spot to mount the switch S1, and carefully drill an appropriate size hole for it. Mark

*Circuit Board Specialists, Post Office Box 969, Pueblo, Colorado 81001. The complete kit of parts including switch and circuit board is priced at $11.00; the circuit board alone is $3.00.

By Robert D. Shriner, WAØUZO, 1740 East 15th Street, Pueblo, Colorado 81001.

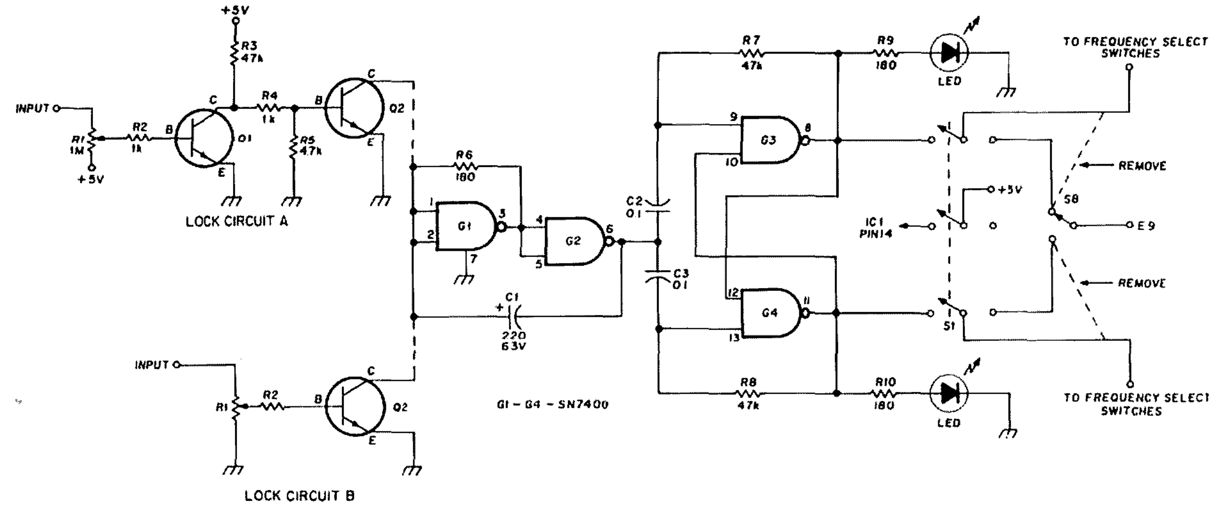



**fig. 1.** Schematic diagram of the channel scanner. The leads to the frequency-select switches provide a high or low voltage that alternately programs the counters in the GLB 400 to the correct frequency. The appropriate lock circuit must be chosen for the receiver used.

your panel with "instant lettering" – scanner *on* and *off*.

The indicator lights should be mounted in such a manner on the front panel so that they will show which set of switches is activated. Drill two holes to fit the light-emitting diodes. After locating switch S8 in your Channelizer, remove the two wires that are connected to the frequency-select switches and connect these to the new switch as shown. Five volts for the scanner is also run through S1 so that in normal operation (not scanning) the LEDs will not be blinking on and off.

This little circuit has added a very valuable feature to an already fine piece of equipment. I do not know how many other synthesizers this will adapt to but I am sure that they exist.

**ham radio**

# improved method for calibrating time-base oscillators

## Two receivers coupled to a common vfo plus an a-m detector and vtvm yield accuracies to better than 2 parts in $10^8$

**Here is a method** that will eliminate the uncertainties of time-base oscillator calibration when listening for a zero beat with WWV or when watching a receiver S-meter. A much more accurate technique is presented for measuring time-base oscillator calibration and stability.

Consider a frequency counter made from a kit and calibrated by ear to WWV. Such a subjective method has the potential for missing errors in frequency calibration over several seconds. There is also a chance for error due to WWV carrier fading. Both of these problems translate directly into TBO calibration inaccuracies.

### calibration method

**Fig. 1** shows a method that virtually eliminates these chances for error. With two high-frequency receivers, a vtvm and some ingenuity, a frequency-counter TBO can be calibrated to an accuracy better than 2 parts in $10^8$ cycles. Note that the two receivers in fig. 1 are coupled to the same vfo (both have a 455-kHz i-f). One of the receivers is tuned to WWV, and the other is tuned to the TBO harmonic that matches the WWV carrier frequency being received. With both receivers using the same vfo, any phase difference between the two i-fs would be attributable to a phase difference between the TBO and WWV.

**Fig. 2** shows the two i-fs fed into an a-m detector. The vtvm is connected to the agc bus and measures the agc voltage developed by the two signals. As long as the two i-f voltages remain constant in amplitude, the only thing that will cause the agc to fluctuate is any phase differences between the TBO and WWV signals. If the two signals vary in phase, fluctuations in the agc voltages will occur.

The WWV signal must be strong and steady for this method to work. My HQ-150 receiver was converted to solid state and the fets in the i-f stage were forced into hard limiting of the carrier by a high rf gain control setting. This principle is identical to fm limiting, where undesirable amplitude modulation of the fm-carrier is clipped. Care was also taken to make sure the TBO did not overdrive the receiver front end. This was done to keep the receiver from radiating signals into the WWV receiver front end. To prevent interaction between the two i-f strips, an IC CA3020 tuned amplifier was placed between the HQ-150 receiver and the a-m detector.*

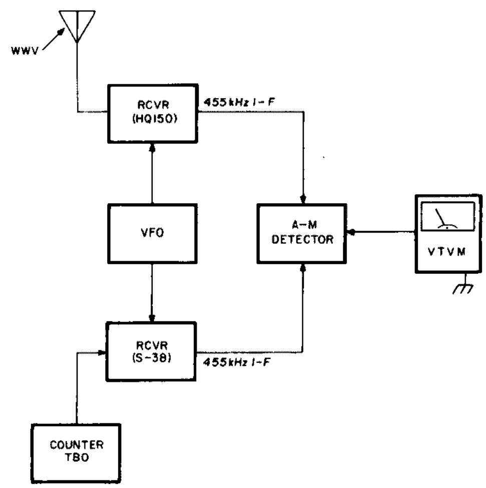



**fig. 1. System block diagram using two receivers and a common vfo. One receiver is tuned to WWV; the other is tuned to the time-base oscillator harmonic that matches the WWV carrier. The S-38 vfo is disabled and the HQ-150 vfo is fed to the S-38 mixer through a CA3020 amplifier circuit (see text).**

Once the receivers had been properly set up, the TBO was calibrated. Adjusting the TBO crystal-oscillator trimmer capacitor caused the vtvm needle fluctuations to speed up or slow down as the TBO became more or less in phase with WWV. Once the slowest possible move-

*The CA3020 amplifier can be found on page 234 of the 1970 RCA *Linear Integrated Circuits Handbook*. My circuit was identical to that of the text with the exception of the substitution of a 455 kHz primary winding from an old i-f transformer in place of the coil shown in the text. This was done to tune the amplifier to 455 kHz.

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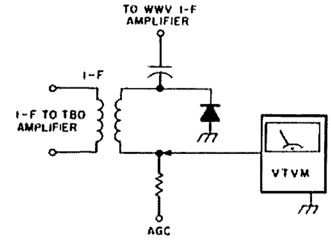



**fig. 2. I-f output of the two receivers is fed to an a-m detector. The vtvm is connected to the agc bus and measures the agc voltage developed by the two signals.**

ment had been observed, the period length of the error was measured.

The vtvm dc-scale level was set for a value that allowed the fluctuations to be observed over a wide enough arc to ensure accurate timing of the cycle. As the needle moved from one end of the scale to the other and back again, the period length of the movement was timed. If the movement took ten seconds and the receiver was tuned to 10 MHz, then the TBO error was computed as one cycle in 100 MHz of WWV; *i.e.*, 10 MHz x 10 seconds = 100 MHz. The error is then of 1 part in $10^8$ cycles.

## accuracy

The accuracy of this method was verified by calibrating the TBO as described above and comparing the results with a laboratory standard, in this case a Gertsch RLF-1 60-kHz WWV receiver-comparator using a comparator frequency of 300 kHz. **Fig. 3** shows the chart-recorder results. In one hour there were 21 cycles of TBO drift. Dividing this drift by (300 kHz x 60 seconds x 60 minutes); *i.e.*, $21/(1.08 \times 10^9)$, yields 1.94 parts of error in $10^8$ cycles. The results confirmed by the Gertsch comparator were almost identical with those computed by the dual-receiver, single vfo method of fig. 1.

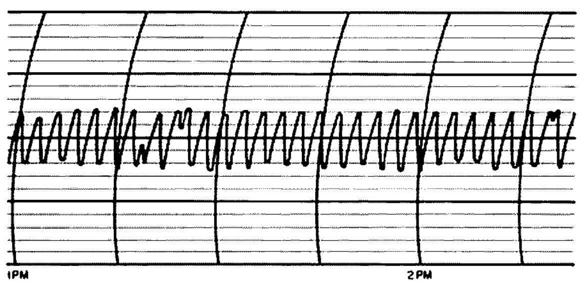



**fig. 3. Calibration results of the measurement method using a Gertsch RLF-1 60-kHz WWV receiver-comparator and a comparator frequency of 300 kHz. About 21 cycles of TBO drift occurred in one hour, yielding 1.9 parts of error in $10^8$ cycles.**

The TBO used for this experiment was enclosed in a crystal oven, and the entire unit was housed in a Styrofoam box to further isolate the TBO from the environment. The procedure described here provides an excellent method for a more accurate calibration of any TBO.

**ham radio**

MICROPROCESSORS

# microcomputer interfacing: the MOV and MVI 8080 instructions

In the preceding column, we indicated that within an 8080 microprocessor chip there exist seven general purpose registers (B, C, D, E, H, L, and the accumulator) each of which operates on eight bits at a time. These registers are used for many varied purposes, *e.g.*, the storage of an 8-bit constant, an 8-bit timing byte, a 16-bit pointer address, or an intermediate result from an arithmetic or logical operation.

By examining the data transfer instructions, MOV D,S, in the 8080 instruction set it can be seen how the general purpose registers may be used. We will assume that there is some data initially present within each register. There are sixty-three different MOV instructions, each of which specifies both the source register, S, of the data and the destination register, D, to which the data is moved. For example, to move data from register E to register B, you would use the instruction MOV B,E which is the Intel mnemonic notation for the operation of moving data to register B from register E. Unfortunately, an 8080-based microcomputer has no way to understand or interpret a mnemonic instruction such as MOV B,E. What is required is the binary representation for MOV B,E, – $01000011_2$. These eight bits, which comprise the instruction code for MOV B,E, can be manipulated digitally, *i.e.*, they can be stored in a semiconductor memory device, transmitted over a data bus, stored in an instruction register within the 8080 chip, and decoded by an instruction decoder into a series of actions that the 8080 performs internally.

**By Jonathan Titus, David G. Larsen, WB4HYJ, and Peter R. Rony.**

Mr. Larsen, Department of Chemistry, and Dr. Rony, Department of Chemical Engineering, are with the Virginia Polytechnic Institute and State University, Blacksburg, Virginia. Mr. Jonathan Titus is President of Tychon, Inc., Blacksburg, Virginia.

How do you convert from the general mnemonic for a move instruction, MOV D,S, to the specific binary instruction code? First, the general form of any MOV instruction is,

| 0 1 | d d d | s s s |
|---|---|---|
| MOV class of instructions | 3-bit binary code for destination register | 3-bit binary code for source register |

and second, each general-purpose register has associated with it a unique 3-bit code.

| register | 3-bit register code |
|---|---|
| B | 000 |
| C | 001 |
| D | 010 |
| E | 011 |
| H | 100 |
| L | 101 |
| * | 110 (to be dis- |
| accumulator (A) | 111 cussed later) |

After the source and destination registers have been selected, insert the respective 3-bit codes into the appropriate places in the general MOV instruction format given above. Some examples are as follows:

| data transfer operation | mnemonic | instruction code |
|---|---|---|
| E→B | MOV B,E | 01 000 011 |
| H→A | MOV A,H | 01 111 100 |
| B→C | MOV C,B | 01 001 000 |
| D→L | MOV L,D | 01 101 010 |
| L→D | MOV D,L | 01 010 101 |
| E→E | MOV E,E | 01 011 011 |

Since binary code is difficult to remember, it is con-

venient to represent the above 8-bit instruction codes in octal code. Thus:

| data transfer operation | mnemonic | octal instruction code |
|---|---|---|
| E → B | MOV B,E | 103 |
| H → A | MOV A,H | 174 |
| B → C | MOV C,B | 110 |
| D → L | MOV L,D | 152 |
| L → D | MOV D,L | 125 |
| E → E | MOV E,E | 133 |

In each case, you have copied data from one register into another. The destination register contains the copy, *the source register remains unchanged.* Notice that the MOV L,D and MOV D,L instructions have the source and destination registers reversed. In the MOV E,E instruction, you have copied the contents of register E back into register E. The net result is that E remains unchanged. This is a valid 8080 instruction, but it has no visible effect and can be called a "do nothing" instruction. Similar "do nothing" instructions exist for the other six general purpose registers.

You may recall that the IN and OUT instructions permit data transfer to occur between the accumulator (register A) and external I/O devices. The MOV D,S instructions offer one means of temporarily storing input and output data bytes elsewhere within the 8080 chip. Besides the IN instruction, how else can data be transferred into the general purpose registers? There are generally two ways of doing so, from the program (for permanent constants), and from memory (for temporary constants, results, data files, etc.).

To get data directly from a program into one of the registers, *immediate* instructions can be used. These are multi-byte instructions that contain the desired data within *the* instruction. The first instruction byte is always the 8080 instruction code; it tells the 8080 chip what to do next. The next one or two instruction bytes contain the actual data. The two-byte *move immediate* instructions,

MVI r
<B2>

permit you to move the data contained in the second byte into the specified register, r. The < B2 > notation means that space must be left in the program for the second instruction byte. The general form of the MVI r instruction is,

| | | |
|---|---|---|
| 0 0 | d d d | 1 1 0 |
| instruction class | 3-bit code for destination register | |

or 0D6 in octal code. Some examples include:

| data transfer operation | mnemonic | octal instruction code |
|---|---|---|
| < B2 > → A | MVI A | 076 |
| | < B2 > | < B2 > |
| < B2 > → B | MVI B | 006 |
| | < B2 > | < B2 > |
| < B2 > → H | MVI H | 046 |
| | < B2 > | < B2 > |

To output the ASCII letter Q to output port 6, you would execute the following program:

| octal instruction code | mnemonic |
|---|---|
| 076 | MVI A |
| 321 (ASCII Q) | 321 |
| 323 | OUT |
| 006 | 006 |

As indicated in the program, you first load the 8-bit ASCII code into the accumulator register; having done so, you then output the accumulator contents to port 6. Data or constants can be moved into individual registers at any point in an 8080 microcomputer program, and as often as needed, simply through the use of MVI r instructions. Remember, each MVI r instruction consists of two instruction bytes.

The transfer of data between memory and the general purpose registers is more complex since *you must clearly and unambiguously specify which one of a possible 65,536 different locations you wish to use for the transfer operation.* This requires a 16-bit address that is stored in the H,L register pair, register H containing the *high* (HI) address byte and register L containing the *low* (LO) address byte. Once you specify these two address bytes, you can readily transfer data back and forth between *the specified memory locations M* and any of the seven general purpose registers. To do so, you use a MOV, D,S instruction, where the 3-bit code for memory location M is 110 (or 6 in octal code).

As an example, if you wish to transfer data from memory location HI = 030 and LO = 123 to register D, you execute the following program:

| octal instruction code | mnemonic | comments |
|---|---|---|
| 056 | MVI L | Load register L with the following LO address byte |
| 123 | 123 | LO address byte |
| 046 | MVI H | Load register H with the following HI address byte |
| 030 | 030 | HI address byte |
| 126 | MOV D,M | Move data from the memory location addressed by register pair H,L to register D |

Remember, whenever you perform an 8080 instruction involving memory location M as "register M," which has a register code of $6_8$, you must specify beforehand the absolute memory address of M in register pair H,L. We shall continue this discussion of the 8080 instruction set in our next column.

**ham radio**

# the ham notebook

## drilling template for integrated circuits

Any amateur scanning the amateur magazines is amazed at the diverse and ambitious construction projects presented. One item that might prevent a large number of amateurs from tackling similar projects is the problem of making printed-circuit boards. Most amateurs lack the necessary equipment and area to fabricate PC boards by the photographic method. The sticky-back foil patterns available are relatively expensive and can result in the board costing more than the components. The remaining technique of resist pen and ink is slow but economical. In the past,

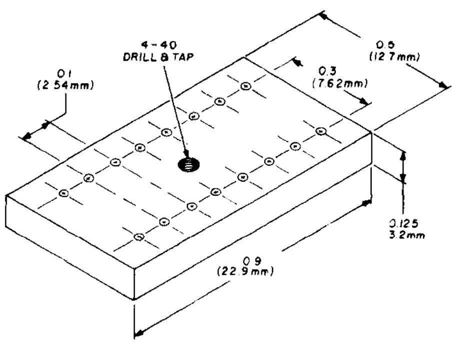



**fig. 1. Dimensions of template block for drilling DIP holes in printed-circuit boards.**

straight sketching on the cleaned copper-clad board was adequate for simple transistor circuits, but the advent of dual-in-line packages has imposed a need for higher accuracy.

One method to fill the gap between foil and resist-pen techniques is shown in **fig. 1**. A 1/8-inch (3mm) thick plate has been precision-drilled for a 16-pin dual-in-line hole pattern. The same plate can be used for 8- or 14-pin layouts. In practice, a single hole is drilled with a no. 32 (3mm) drill and the plate is

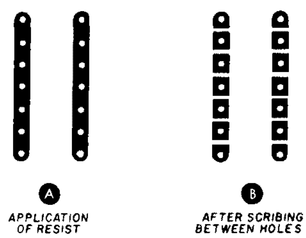



**fig. 2. Applying resist pattern for DIP IC packages, (A) application of resist; (B) after scribing between holes.**

clamped to the foil side of the board with a single 4-40 (M3) screw through the hole. With the plate in place, the drilling of the required hole pattern is simple and rapid. The result is an accurate hole pattern with evenly spaced holes.

Next, the pads must be marked with a resist pen. This is easily accomplished by first marking each line of holes with a continuous stroke as shown in **fig. 2A**. Then a scribe is used to separate and isolate each hole pad as shown in **fig. 2B**. Using this method, a clock driver circuit using four 14-pin packages was drilled, laid out, etched, and completed in less than two hours.

The drilling template can be made by any amateur or professional machinist. We were able to get a local machinist to make them for $4.00 each from brass and $6.00 each from stainless steel. If you have trouble getting one made up, we would be glad to sell them at cost to any amateur.

John M. Franke, WA4WDL
Norman V. Cohen, WB4LJM

## heterodyne crystal switching in Heathkit SB-series equipment

Heterodyne crystal frequency changing has come into some use due to the desire by many amateurs to communicate with Oscar and MARS stations. Rather than permanently replacing the crystal (which seems permanent) or using a relay to switch the crystal (which isn't really necessary), I devised the following means.

A small Z-shaped aluminum bracket was secured to the chassis between tubes V-10/11 and V19 by two 4-40 (M3) nuts and screws which replaced the existing ones. A crystal socket and a spdt switch are mounted on the bracket and wires as shown in **fig. 3**. This enables *any* heterodyne crystal to be

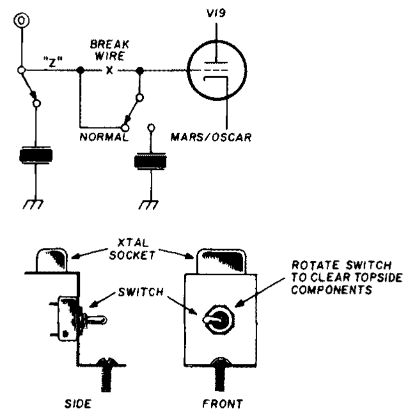



**fig. 3. System for improved heterodyne crystal switching in the Heath SB-series equipment.**

changed easily. Nothing is lost or permanently changed, just added and easily removed. I used a socket which accepts FT-243 and HC-17/U holders since pins can be soldered to HC-6/U holders to fit that socket as well.

Paul Pagel, K1KXA

## Drake R4-C modification

At low listening levels, when using headphones or speaker with the Drake R4-C receiver, the audio is noticeably contaminated with power supply ripple. This may be corrected by a remarkably simple modification which will become obvious after examination of the chassis layout.

In order to connect the diode and regulator board located near the rear of the chassis with the filter capacitors located near the front of the chassis, connecting wires have been run in a circuitous manner alongside the audio cables and through the wiring harness. Return currents to the power transformer center taps travel through the chassis underneath the audio board ground connections.

To re-route these ground currents, simply clip the red-yellow transformer lead from the grounded tie-point at the rear of the chassis and, by splicing in another piece of wire, connect it to a ground lug of filter capacitor C163. In a similar manner, move the blue-yellow transformer lead to a ground lug of filter capacitor C166, and dress both wires along the inner edge of the chassis. Next, locate the white-purple lead which goes into the wiring harness and emerges again to connect with capacitor C166 Δ, clip one end of this wire from the base of regulator transistor Q2, and clip the other end of the wire where it is fastened to the capacitor. Replace this wire with another and dress it with the two wires already installed and dressed along the inside edge of the chassis as previously described. This procedure should be repeated with the white-orange lead to capacitor C166D, and the white-yellow lead to C163□. Additional capacitance across filter capacitor C166 helps to a small degree, but the major improvement is obtained simply by moving the wires as described.

George R. Bailey, WA3HLT

## Heath SB-102 modifications

I have previously written about R955, a 100k resistor, being underrated. It avalanches down in value and burns up. Replace R955 with a 1- or 2-watt resistor (page 58, June, 1975, *ham radio*). There is at least one other resistor, R103, which has *similarly* burned up. Replace it with a 1- or 2-watt resistor as well.

In both of my Heath SB-102s I have encountered another problem which is especially bad in the CW mode. After transmitting for a couple of minutes and then reverting to the receive condition, the S-meter drops nearly to zero for two or three seconds and then quickly builds up to a high value (S9+20 dB or more), which desensitizes the receiver to a weak incoming signal for about 30 to 60 seconds. There is an easy top-of-the-board fix for this problem, however.

Without going into all the detail, tube V2 is acting as a noise generator for the short time that screen voltage remains on the tube after reverting to the receive mode. I found an easy way to cut this stage off promptly when reverting to the receive condition, without unduly upsetting the ALC action of the stage. After much "cut and try" of values, the following was adopted as a satisfactory solution to the problem.

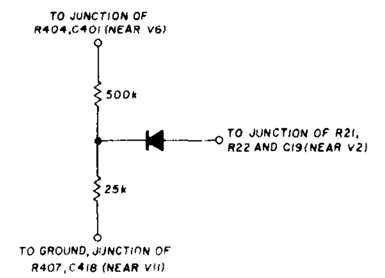



**fig. 4. V2 will be cut off completely using this addition to the bias.**

Locate R404 (near V6) where it joins C401. Attach one end of a 500k, ½-watt resistor to this point. Connect the other end of the new 500k resistor to the cathode end of a diode *and* to one end of a 25k, ½-watt resistor. The other end of the 25k resistor goes to ground, which can be conveniently picked up at the junction point of R407 and C418 (near V11). The other end of the diode goes to the junction point of R21 and R22. The schematic diagram is shown in fig. 4.

Lowell White, W2CNQ

## suppression of rf interference in telephones

Contemporary telephone design often fails to take into account interference from a potentially large number of rf energy sources. Varistors are used in telephones as high-level audio limiting devices, but they also detect and rectify rf, making each telephone behave somewhat like a crystal receiving set in the presence of strong rf fields.

Several telephone manufacturers have designed and developed telephones capable of satisfactory operation in the presence of strong rf fields, but these special sets are not normally stocked by most local telephone company offices. Therefore, it is desirable, if not essential, to find another means of filtering rf energy from ordinary telephones. Fortunately, a simple modification to the telephone renders the set immune to rf interference and makes complete rf suppression possible.

An inductor, designated 1542-A, consists of two rf chokes and serves as an effective rf filter at broadcast frequencies. Moreover, 1542-A inductors are readily available at most telephone company offices and are often installed in home telephones to suppress rf from powerful local broadcast stations. Unfortunately, at some amateur frequecies, the rf chokes of 1542-A inductors becomes highly reactive and may burn up in the presence of very strong rf fields. Although such an occurence is unusual, it can happen, and may be easily prevented in the following manner.

For Bell Telephone series-500 telephone sets:

1. Place a 1542-A inductor in series with the telephone set.

2. Connect a 0.01 $\mu$F capacitor across the receiver element.

3. Connect a 0.01 $\mu$F capacitor between terminal RR and terminal C of the telephone network.

4. Connect a 0.01 $\mu$F capacitor between terminal GN and terminal R of the telephone network.

For General Telephone series-80 telephone sets:

1. Place a 1542-A inductor in series with the telephone set.

2. Connect a 0.01 $\mu$F capacitor across the receiver element.

3. Connect a 0.01 $\mu$F capacitor between terminal 4 and terminal 23 of the telephone network.

4. Connect a 0.01 $\mu$F capacitor between terminal 1 and terminal 23 of the telephone network.

These modifications render the telephone immune to interference from rf energy and disable it as a crystal set. The 1542-A inductor limits the rf passed to the telephone, and the capacitors bypass rf energy.

When building or installing a phone patch, it is desirable to install a 1542-A filter in the telephone line to the patch. If a filter is not available, two 2.5 mH rf chokes can be installed in series with the telephone line to the patch, one choke in each side of the line. They should be mounted inside the patch cabinet or box, and a 0.01 $\mu$F capacitor connected across the telephone line on the patch side of the rf chokes. The microphone side of the patch should also be decoupled by a 2.5 mH rf choke placed in series with the microphone lead inside the phone patch. A 470 pF bypass capacitor should be connected between the microphone side of the rf choke and ground. With this simple modification, telephone patching is immune to rf interference.

Rf detection by phone patches and telephone sets can also be a good indication of a poor ground or a poor antenna system at the rf source. It is therefore desirable to have a good antenna and ground system at the transmitter for radiation efficiency as well as for reducing stray rf energy.

Ken Anderson, K7LDZ

## S-meter for Genave transceivers

Since Genave transceivers do not have tuning meters or S-meters to assist in tuning weak stations, I installed the simple circuit of fig. 5 in my GTX-200 to take care of this omission. I drilled a hole for a phone jack in the rear panel and installed the rest of the components by short leads on the bottom of the board. I have used a Heath IM-104 fet vom voltohmmeter and have also used a 25 micro-amp meter, direct. On my Genave, IC101 is the same as an RCA 3065E.

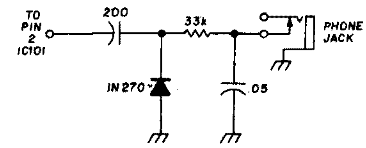



**fig. 5. Simple circuit for adding an S-meter to the Genave two-meter transceiver. Meter is connected to the phono jack.**

Larry M. Chrisman, K9OXX

## low-cost, two-meter mobile antenna

Have you ever wanted a simple two-meter mobile antenna that could be used in a hurry, that didn't have to be permanently mounted, that didn't cost much, and didn'trequire any hole drilling? I was in that category a few months ago — on a Sunday, as I recall. The stores were closed, but I wanted to go mobile with my TR-22 for a short trip and work some repeaters along the way. Here's how I solved the problem (with help from my wife).

I used a large stationery spring-clip, the kind with wide, flat clamping jaws and a pair of tabs or "ears" with holes in them. The holes looked to be the right size to accept the shell of an auto-type antenna connector that I happened to have in the junk box. I reamed the hole just a bit to accept the connector body, and bent the connector tabs over the edge of the clamp "ears" to secure the two together. After this was done, I attached a short length of RG-58/U coax to the assembly; shield braid to the clamp-shell and center conductor to the "hot" pin on the connector. Now for the antenna element.

I used the mating auto-antenna plug to mount a 19-inch (48.3cm) piece of music wire soldered to the tip. For insulation and security, I filled the top of the plug surrounding the wire with some five-minute epoxy and let it cure. Presto! The antenna was done, and it took only about one-half hour to build.

Results were just great. I clipped the makeshift antenna to the rain gutter molding on the mobile, and bent the tab holding the antenna just enough so that the antenna was vertical. The coax was led through the partly-open window to the TR-22 on the seat next to me. Just for fun, I put an swr bridge in the circuit and was surprised — and pleased — to find that it was less than about 1.3:1. Until I bought my new 5/8-wave magnetic mount antenna, this little gutter-clamp job did the trick. Neither the electrons, nor the amateurs I talked to, knew the difference.

Jim Gray, W2EUQ

## IC230 modification

The ICOM model IC-230 transceiver has a push-button selector switch on the front panel labelled "megacycle switch." Also a slide switch on the top right corner, labelled A/B, provides ±600-kHz offset for repeater operation. When switching from 146 to 147 MHz or vice versa, the A/B switch must also be switched. Failure to do so, or forgetting what position it should be in, could put you outside the band. I have incorporated a minor modification to my IC230 to allow proper offset under all conditions when the A/B switch is left in position A (fig. 6). For special applications, opposite shift may be selected by switching the A/B switch to position B.

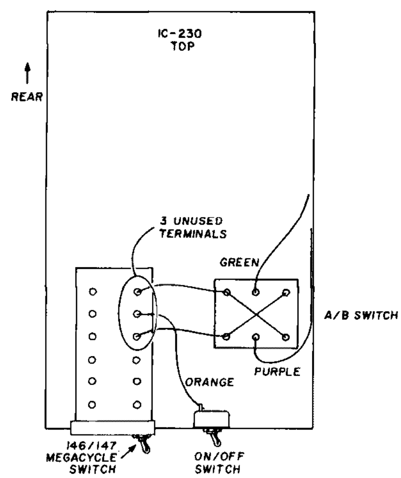
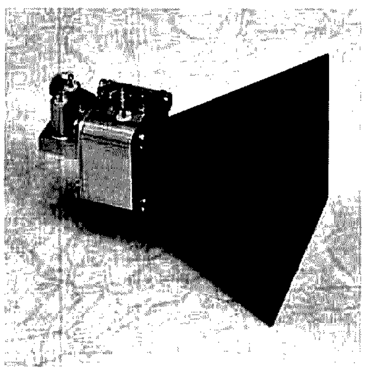



**fig. 6. IC-230 modification to avoid out-of-band operation.**

Bill Theeringer, W8PEY

# NEW products

## Amateur 10-GHz Gunnplexer



Microwave Associates has recently announced a new solid-state transceiver designed especially for the Amateur 10-GHz band which consists of a varactor-tuned, frequency-modulated Gunn-diode oscillator and Schottky mixer diode. The rear portion of the unit pictured above consists of a Gunn-diode oscillator which directly converts dc (+10 Vdc nominal) to 20 mW of rf energy. The oscillator is preset to 10.250 GHz, the center of the Amateur band, unless otherwise specified. Mechanical tuning allows the center frequency to be tuned ±100 MHz; the built-in varactor is used for frequency modulating the device and will deviate the center frequency up to 60 MHz, depending upon the applied tuning voltage.

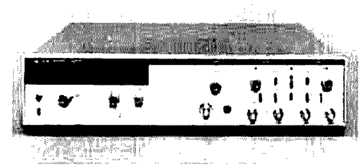
The receiver section of the *Gunnplexer* consists of a low-noise Schottky mixer diode which has a noise figure of about 12 dB when used with a low-noise i-f strip. A small amount of power from the Gunn oscillator (about 0.5 mW) is applied to the mixer diode through a ferrite circulator integrated into the waveguide mount. The circulator also isolates the receiver and transmitter functions of the module. A horn antenna with 17 dB gain is available as an accessory.

Since the Gunn oscillator provides both transmit power and mixer injection, the setup for radio communications is similar to that used with klystron-based polaplexers (see page 40, this issue), *i.e.*, the center frequencies of the *Gunnplexers* used at each end of a communications link must be offset by the i-f frequency, typically 30 MHz. Fm broadcast receivers can also be used for the i-f system, with afc applied to the *Gunnplexer* tuning varactor to minimize frequency drift (which is −350 kHz/°C maximum without afc).

Maximum range with the 20 mW Microwave Associates *Gunnplexer* depends on a number of factors including required fm signal-to-noise ratio, fm deviation, i-f bandwidth, and antenna gain. However, assuming a 10-dB signal-to-noise ratio (the threshold of speech in an fm system), a line-of-sight path between stations, and the accessory 17-dB horn antennas, maximum range is about 25 miles with a 240-kHz i-f bandwidth (that used in fm broadcast receivers). If the i-f bandwidth is decreased to 50 kHz, maximum range is about 55 miles.

The *Gunnplexers* can be used for two-way communications, as a link between repeater sites, security alarms, or fm doppler radar systems. A separate power supply, fm modulator, and receiver must be provided by the user. The MA87127 *Gunnplexer* is priced at $85. Also available is the MA87108 which consists of the Gunn oscillator and tuning diode module ($60). A *Gunnplexer* with the accessory 17-dB antenna is priced at $108. Two complete transceivers with horn antennas, MA87141, are $185. For more information, write to Microwave Associates, Inc., Northwest Industrial Park, Burlington, Massachusetts 01803.

## H-P frequency counter extends range to 1300 MHz



Frequency measurements to 1300 MHz covering mobile communications bands, vhf and uhf television, am and fm broadcast, TACAN and DME frequencies are now possible with Hewlett-Packard's Model 5328A Universal Counter using a new module. Called Option 031, the module adds five features to the counter's capabilities:

1) Extends frequency range to 1300 MHz

2) Provides 20 millivolt sensitivity for measuring low-level signals

3) Adds automatic gain control (AGC) for protection against overload

4) Includes a fused 50 ohm input channel

5) Provides accessory power for a 20 dB preamplifier for more sensitivity.

The basic Model 5328A with no options makes frequency measurements to 100 MHz, and single-shot time interval measurements to 100 nanoseconds reso-



More Details? CHECK—OFF Page 126

lution. Time interval averaging increases resolution to 10 picoseconds for repetitive events. It also measures period, period average and frequency ratio, and will totalize and scale inputs. Frequency measurement sensitivity is 25 millivolts rms to 40 MHz and 50 millivolts rms to 100 MHz. With its eight-digit display and a standard crystal time base, the basic Model 5328A sells for $1300. U.S. price of Option 031 is $600, and delivery is 4 weeks.

For additional information write to Hewlett-Packard Company, 1501 Page Mill Road, Palo Alto, California 94304, or use *check-off* on page 126.

## circuit-stik/centron catalog

Circuit-Stik and its sister company, Centron Engineering, announce the release of their new condensed catalog no. 503, merging three distinct product areas of the two companies: Circuit-Stik's unique sub-elements for making instant circuit boards, pre-etched GP (general purpose) prototype and wire wrap socket boards, and Centron Engineering's printed circuit drafting aids for making circuit board master artworks.

Each area features new products and detailed information for many applications, assembly ease and time-saving features. A *How to Use* section with technical specifications is included. Several new microprocessor oriented GP boards, as well as new card cages and accessories, are featured in the GP Board section. New IC DIP continuous patterns are offered in the drafting aid area.

For further information, contact: Circuit-Stik/Centron Engineering, Inc., 24015 Garnier Street, Torrance, California 90505.

## variable speed tape transport for microprocessors

The Triple I Division of The Economy Company of Oklahoma City is now offering a new low-cost variable-speed model to its line of electronic cassette tape transports. Features of the variable speed model include: four-

motor control, remote control capabilities, fast start-stop, less than 30-second rewind, ac or battery-operated and variable speeds (0.4 to 10 IPS). Nominal power requirements are 7.0 volts dc at 600 mA.

In addition to use in microprocessors, the unit has applications in data recording/logging/storage, programming, instrumentation, industrial controls, data duplicating, security/automatic warning systems, testing apparatus, audio-visual education, hi-fi, and other general applications.

Four separate motors control take-up, rewind, play or record, and head engagement. The separate motors allow the most complex tape deck function to be accomplished by remote control. Flutter, wow and jitter are minimal because the capstan drive motor has only one job – moving the capstan.

Control Boards for the Phi-Decks, which are TTL, DTL, and CMOS compatible, contain all the circuitry for proper control of the Phi-Deck tape transport.

Options such as EOT/BOT sensing, record/play, read/write, electronics, cassette-in-place sensing, and others are available.

For further information, write or call Triple I, a division of The Economy Company, 1901 North Walnut, Box 25308, Oklahoma City, Oklahoma 73125; telephone (405) 528-8444, Ext. 71 or 79, or use *check-off* on page 126.

## four-channel hi-fi decoder

National Semiconductor Corporation and Tate Audio Limited have combined efforts to produce a kit of three types of integrated circuits that will accurately reproduce quadraphonic four-channel high-fidelity sound from phonograph records. Quad hi-fi, which can reproduce sound with concert-hall realism, is expected to become a standard format of the recording industry, and could eventually replace stereophonic sound formats.

Tate designed its system to decode CBS SQ-type program material from phonograph records and tape cassettes after making an exhaustive analysis of the theories and practicality of all techniques for reproducing four channels of sound. The SQ technique uses standard,



More Details? CHECK—OFF Page 126

existing pick-up components that are compatible with monaural and stereophonic formats. In addition, the SQ format is compatible with stereophonic broadcasting methods to make use of a broad range of records and tapes already produced by CBS and other major companies.

Separation of channels in any direction approaches 40 dB across the entire audio bandwidth, while the signal-to-noise ratio is 70 dB. Total harmonic distortion is only 0.05 per cent. These parameters meet the most demanding specifications of high-fidelity sound technology.

Although the Tate system is sophisticated in concept, its comparatively low number of external components permits a circuit board size of only 1.5x3.25 inches (36x80mm).

Additional information may be obtained by contacting Roy L. Twitty, National Semiconductor Corporation, 2900 Semiconductor Drive, Santa Clara, California 95051; telephone (408) 737-5287; or Wes Ruggles, Tate Audio Limited; telephone (213) 822-3189; or use *check-off* on page 126.

## microcircuit computes true rms value

The first genuinely low-cost *true rms* converter with guaranteed accuracy, crest factor, and frequency response is now available from National Semiconductor. Known as the model LH0091, the new converter will compute the root-mean-square value of virtually any combination of ac or dc input signal from dc to 2 MHz. The device is ideal for use in digital voltmeters, in digital multimeters, in measuring audio signals or noise levels, in making harmonic or vibrational analyses, and in power measurement and control. An uncommitted amplifier is provided for filtering, gain, or high-crest-factor configuration.

The *LH0091* is manufactured in two kinds of 16-lead dual in-line packages. Available in both commercial and military temperature range, the device is priced between $22.50 and $44.00, depending on the temperature range. For more information write National Semiconductor, 2900 Semiconductor Drive, Santa Clara, California 95051.

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# radio magazine

solid-state
microwave
rf generators

# ham radio

*magazine*

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**volume 10, number 4**



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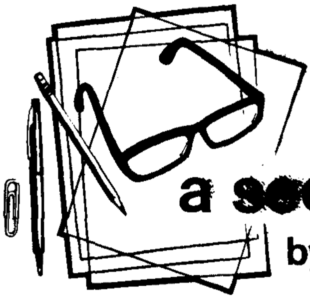


# a second look

**by Jim Fisk**

A new era in community education may well be the outcome if a petition filed recently with the FCC receives favorable action. Based largely on experience gained with vhf repeaters, and more specifically, amateur television repeaters, The Center for Advanced Study in Education of the City University of New York has filed a petition for the establishment of a new community educational radio service to be known as *Communicasting*. Called communicasting because it embodies elements of both communications and broadcasting, the new service would use television channels 70 through 83 for co-channel, multilateral video and audio communications. Using inexpensive terminals in homes, schools, community centers, libraries and hospitals, the system would tie the community together in an interactive educational network. The petition is co-sponsored by the Communicasting Association of America, a non-profit organization headed by W2KPQ which is dedicated to using the radio spectrum for multilateral educational and scientific communications.

If Communicasting is approved, any individual at home would be able to receive the transmissions on one of the unused vhf channels of his television set. The low-power signals from the remote terminals would be transmitted to a translator where they would be re-transmitted on one of the uhf TV channels. The antennas would be high enough to cover the entire community. It is an idea that can effectively and inexpensively implement the concept of "Communicate instead of Commute" by providing electronic classrooms, forums, and lecture halls.

One of the first examples of Communicasting was established on the amateur two-meter band in 1955 when the Albany Medical College started a novel form of post-graduate education: Two or three members of the faculty discussed a medical topic while in direct radio communication with doctors located in outlying hospitals. The conference network consisted of a high-power transmitter at Albany Medical College and lower-powered units at twenty-one hospitals throughout eastern New York and western Massachusetts. The system is functioning to this day.

The principal of Communicasting was further demonstrated on the MARS frequencies in 1958-1960 where it was used for on-the-air scientific and educational forums. Today it is being used for weekly technical nets which are transmitted through a vhf repeater in the New York metropolitan area. Further experiments will be conducted in future months on the uhf television repeater recently unveiled by the Long Island Mobile Amateur Radio Club.

Many educators have recognized the potential of interactive radio and television in traditional classroom activities as well as in continuing and extension programs, and homebound education. At the City University of New York, the Center for Advanced Study in Education and the Institute for Research and Development in Occupational Education have been actively developing courses of study for electronic classrooms, studying the most effective way of delivering the curriculum, and assessing the coverage available with direct transmission and with repeaters. Their research will have a direct bearing on initial efforts to demonstrate a working system in New York State during the 1977-1978 academic year.

The FCC recognized the need for an "Educational Radio Service" in 1963 when they established the Instructional Television Fixed Service (ITFS) in the 2.5-2.69 GHz band. However, since equipment for these frequencies is up to a hundred times more expensive than equipment for the uhf bands, the use of ITFS is effectively limited to large, wealthy institutions who can afford the equipment. The proposed Communicasting network would put it within the reach of everyone.

As Dr. Lee Cohen of the City University of New York said recently, "At present we are in the stone age of multilateral education and scientific communication by radio. We look forward to the day when the FCC will allocate a band of frequencies wherein professional and educational groups could organize radio forums . . . This could eliminate the problems of time and distance in getting some of our foremost minds to communicate by radio, thereby educating a listening/viewing audience."

If you are interested in supporting this worthwhile proposal, or would like to know more about it, write to Ed Piller, W2KPQ, Communicasting Association of America, Inc., 80 Birchwood Park Drive, Syosset, New York 11791.

**Jim Fisk, W1HR**
**editor-in-chief**

# presstop

FCC'S PROPOSED AMPLIFIER BAN would prohibit the marketing of external RF amplifiers capable of operation from 24 through 35 MHz. In its February 18 Public Notice, the Commission specified its concern with the so-called "broad-band linears," which replaced "business-band" linears in the marketplace after those had been banned two and a half years ago, in February, 1975.

In Limiting Its Proposed Ban to 24-35 MHz the FCC also made note that though the Amateur service would be affected Amateurs would still be permitted to build 10-meter amplifiers or modify commercial units to cover 10 meters for their own use, while they "would respect the intent of this regulation and not supply these devices to non-Amateurs." Such construction by individual Amateurs would be limited to a single unit of a given model.

Some Specific Areas the FCC would like Amateurs to address in their comments are:

Any Further Requirements which may be necessary to prevent the use of illegal amplification devices;

Practicality Of Such a prohibition and possible techniques which could be used to produce such an amplifier;

Problems Associated With preventing the few unscrupulous manufacturers from including such features as accessible wiring which could be cut to provide for operation on the prohibited frequencies; and

Controls Which Could provide for operation on these frequencies, or any other concepts which could be used to circumvent this prohibition.

Comments On This Docket, 21116, are due May 25; Reply Comments must be submitted by June 6.

Amateur Transmitters and amplifiers would both require type acceptance under the terms of Docket 21117. In this NPRM the FCC pointed out that most current Amateur equipment is commercial and some of it is capable of operation on CB frequencies, but type acceptance could help control that capability. Furthermore, though Amateurs bear the basic responsibility for the performance of their equipment, type acceptance would be a means by which that responsibility could be shared by the makers.

Specific Exemptions from the type acceptance requirement for equipment built or modified by Amateurs for their own use were also proposed, as were provisions for type acceptance of kit-built designs. Comments on Docket 21117 are also due May 25 with Reply Comments June 6.

In Both Of These far-reaching Notices of Proposed Rule Making the Commissioners have left the door open fot workable alternative-solutions to the problem of non-Amateur use of Amateur equipment on the CB bands. Three months should provide enough time to find some such solutions. Let's hope so!

SECONDARY AMATEUR STATION LICENSES could be abolished entirely or a moratorium imposed on processing Amateur applications other than those from new, upgrading or renewing Amateurs if the tide of multiple applications presently arriving at Gettysburg isn't stemmed. Since license fees were abolished in January an increasing number of new applicants and the newly eligible 1x2 seekers — now threaten to bury the limited portion of the Gettysburg facility devoted to processing Amateur applications.

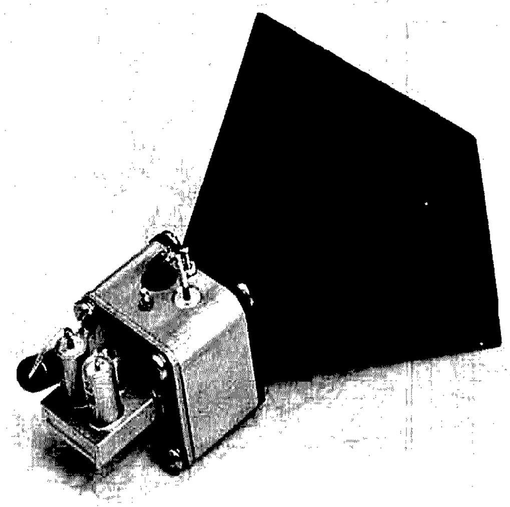
FCC'S NEW NOVICE EXAM has a circuit diagram error in the Ohm's Law problem which makes it impossible to answer as presented. Gettysburg has been advised to give all Novice applicants credit for that question whether it is answered or not — Novice training instructors please note.

AMSAT AND ARRL HAVE SIGNED an agreement in which the League will provide major assistance to the on-going Amateur space program. The League also provided two technicians — WA1JLD and WA1JZC — to bolster the AMSAT effort on the AO-D spacecraft.

The AO-D Satellite (OSCAR 8) is now scheduled for a November 15th launch. It'll carry 145-28 and 145-435 MHz transponders; and, with its 500-mile high orbit, will be even easier to access — though for shorter periods — than the present Amateur satellites, OSCARS 6 and 7.

"AMATEUR RADIO...IN THE PUBLIC INTEREST" is a very attractive report on Amateur Radio in 1976 published by the ARRL for use in presenting the Amateur service to public officials and the media. Copies for PR use are available on request, but do specify with your request how and where they'll be used.

RF POLLUTION is being studied by the Environmental Protection Agency. The EPA says it's concerned about environmental exposure arising from the ever-increasing number of RF sources, expects its extensive monitoring survey to lead to significant data within the next 18-24 months. They've already determined that "significant portions of the population are exposed to 0.1-1 microwatt/square centimeter range" radiation from the 55 to 1000-MHz part of the spectrum.



# solid-state microwave rf generators

A discussion of microwave diode oscillators, including a description of the Gunnplexer — a complete solid-state transceiver for the amateur 10-GHz band

Of the many interesting trends that have developed in radio, perhaps none has been more spectacular than that toward the higher frequencies. No doubt many readers will remember when the term "high frequencies" was used to distinguish 100 kHz from audio, and when the "ultra highs" meant anything above 10 meters. Although most of the early developments in wireless around the turn of the century were accomplished on radio frequencies below 500 kHz, it's interesting to note that Heinrich Hertz's first experiments with wireless transmission in 1887 were conducted near 60 MHz; he later extended them to 500 MHz. In the first decade following Hertz's discovery the frequency frontier was quickly pushed to 75000 MHz. Marconi, in fact, used vhf in many of his early demonstrations, but quickly switched to the lower frequencies when he recognized that greater distances could be covered with the simple spark equipment then in use.

About the same time amateurs were opening up the "short waves" above 1500 kHz with their famous Transatlantic tests of the early 1920s, researchers E. F. Nichols and J. D. Tear had succeeded in producing radio waves as short as 0.22 mm (135 GHz). In 1923 Madame

By **James R. Fisk, W1HR,** *ham radio magazine,* Greenville, New Hampshire 03048

Glagowela-Arkadiewa working in Russia extended the frequency limit to more than 35-million MHz.[1] In all of these experiments the microwave energy appeared as a harmonic component of high-energy spark discharges, and the power level was very low – perhaps microscopic would be a better description.

Soon after the invention of the three-element vacuum tube, work was started toward extending its range into the higher frequencies but there were many difficulties to be solved. Since the early tubes were built around techniques borrowed from the electric lamp industry, they were ill suited for the job at best. Extensions in frequency closely followed improvements in vacuum-tube manufacturing methods, but it wasn't too long before researchers were faced with another problem: electron transit time – the finite time it takes an electron to cross the tube. At the high frequencies the transit time (about a billionth of a second) is short compared to one complete rf cycle, so the electrons can follow the rf voltage fluctuations on the grid. At the very high frequencies, however, the oscillations are so rapid that the voltage on the grid may go through several complete cycles while the electron travels across the tube, and the grid voltage cannot impose its signal pattern on the electron flow. The regenerative vacuum-tube oscillator could be made to work up to 150 or 200 MHz with tuned Lecher lines, but that was about the limit, even with specially designed tubes (two amateurs, Robert Kruse and Boyd Phelps, extended this to nearly 750 MHz in 1927, but that's getting ahead of the story).

In 1920 two German engineers, H. Barkhausen and K. Kurz, found that if the grid of a vacuum triode was biased positively with respect to the plate, they could produce rf output at a wavelength of 43 cm (697 MHz).[2] With this arrangement the highly positive grid accelerates the electrons from the cathode at high speed

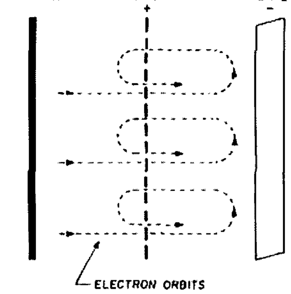
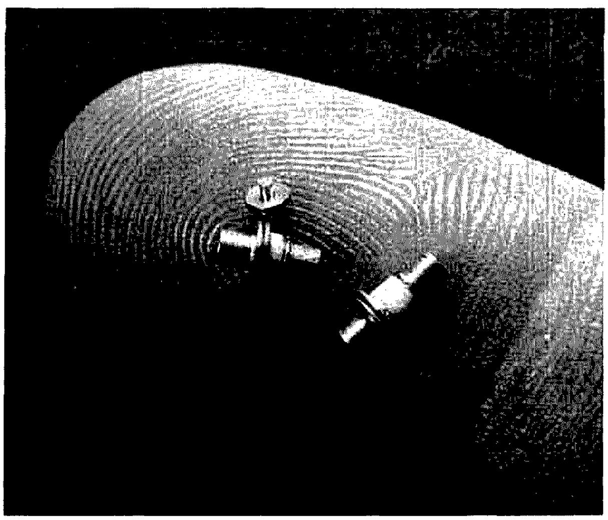



fig. 1. Mechanism of Barkhausen-Kurz oscillators, early method of obtaining uhf rf signals. The positive grid accelerates the electrons from the filament – most strike the grid and give up their energy in heat, but others pass through the openings in the grid, and are repelled by the negatively-charged plate back toward the grid. The electrons continue to orbit between the grid and plate, as shown here. The feeble output is coupled out through a tuned plate line.

(fig. 1) – some hit the grid and give up their energy as heat, but others pass through the openings in the grid only to be repelled back toward the grid by the negatively-charged plate. When the electrode voltages are properly adjusted, the electrons continue to gyrate between the grid and plate at a very high frequency. Barkhausen and Kurz further reported that the frequency of oscillations was dependent upon the applied voltages, with little regard for the external tuned circuit. This phenomenon created quite a stir among researchers, but American engineers were hard pressed to duplicate their results – because of patent fights there was an embargo on foreign vacuum tubes, and the internal construction of American tubes didn't support this mode of operation (a cylindrical, coaxial plate and grid structure was required and American tubes used a flat, sandwich type construction).

Small size of microwave power diodes is misleading – diodes the same size as those shown here are capable of providing CW power outputs of 500 mW or more at 10 GHz.



Eventually the Americans were able to "acquire" some suitable tubes from military sources who apparently weren't affected by domestic trade embargos, and the same oscillations were observed. In 1922 E. Gill and J. Morrell found that if the element voltages were very carefully adjusted, the frequency of oscillation could be controlled by tuned Lecher lines. By suitable modification of the electrodes, frequencies of about 6000 MHz were produced by Kohl in Germany as early as 1928. This is the same rf source used at 600 MHz by Yagi and Uda in 1928 when they were developing the parasitic array. Two years later Dr. Esau used Barkhausen-Kurz transmitters and receivers for full duplex operation across the English channel on 1670 MHz.[3]

The efficiency of the Barkhausen-Kurz oscillator was very low because most of the oscillating electrons were intercepted by the grid – which often ran white hot. In 1921 A. W. Hull proposed a solution to the problem: The magnetron, a device which didn't *require* a grid; the electrons were kept in a circular orbit around the cathode by an external magnetic field (fig. 2). The original design received considerable modification, notably by Yagi and Okabe in Japan who split the anode

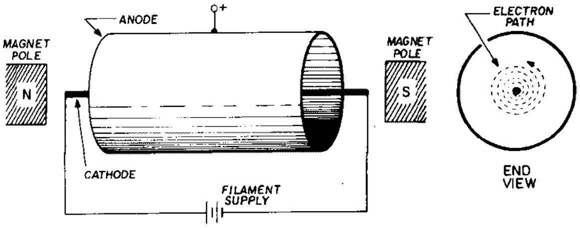



**fig. 2. Basic magnetron. The electrons emitted by the cathode do not reach the positively-charged anode if the magnetic field is strong enough. If the magnetic field is properly adjusted, the electrons go into a spiral orbit around the cathode (end view, right), with the turns oscillating at a rate determined by an external tuned circuit. By 1930 devices such as this were being used to generate small amounts of microwave energy at wavelengths of 3cm (10 GHz).**

into two or more parts and increased both frequency and power output. Although the magnetron offered considerable promise, it ran into some of the same difficulties as other devices – the dimensions of the component parts became so small at microwave frequencies that it was difficult to dissipate heat in the small space which was available. Nevertheless, by 1936 C. Cleeton and N. Williams at the University of Michigan were operating a magnetron at 50 GHz with very limited power output.

Obviously, before microwaves could be put into practical use, some way had to be found to generate useful amounts of power, but researchers were stymied on two fronts: the electrons moved too slowly, and the electrodes of the tube had to be as small as possible because they formed the capacitance of the resonant circuit. For the generation of liberal amounts of rf energy, however, the electrodes had to be as large as possible to dissipate the heat of electron bombardment. Thus, with existing devices, the requirements of microwave rf and high power could not be fulfilled simultaneously. In 1935 Dr. A. Arsenjewa-Heil and Dr. Oskar Heil proposed a unique solution to this mutual incompatibility. Why not, they asked, use the finite transit time of electrons to control the electron stream and derive the oscillation energy directly from the electron stream? In their proposed design the electrons didn't hit the electrodes at all, so the heating problem was completely avoided. Although the Heils were the first to describe the principle of velocity modulation, others had been thinking along the same lines, including Dr. William W. Hansen at Stanford University. Based on Hansen's calculations, Russell and Sigurd Varian put these ideas to work in 1937 when they built the first klystron, a device which uses transit time and the deceleration of bunched electrons across a vacuum gap to generate rf power (fig. 3).

In many respects the development of modern, solid-state microwave devices parallels advances in vacuum-tube technology in the 1920s and 1930s. Transistors suffer from electron drift-time problems at high frequencies, too, and it wasn't too many years ago that amateurs were hoping for a low-cost transistor that they could use successfully on 3.5 MHz. Vhf and uhf transistors are now commonplace, but only because the manufacturers have found ways of making the active region of the transistor wafer thin enough that electron transit time doesn't cause problems.* At microwaves, however, transit time is still the limiting factor, and this is where the analogy between microwave vacuum-tube and semiconductor development comes in: Like the Barkhausen-Kurz oscillator, magnetron, and klystron that preceded them, the operation of Gunn devices, avalanche diodes, and other solid-state microwave rf sources is also based on electron transit time.

In 1953 W. T. Read of Bell Labs proposed a multilayer diode for generating microwave power.[5] Read suggested that the finite delay between an applied rf voltage and the current generated by avalanche breakdown, and the subsequent drift of the generated carriers through the depletion layer of the diode junction would lead to negative resistance at microwave frequencies. The multilayer diode proposed by Read was very difficult to build, but in 1965 R. L. Johnston and his colleagues at Bell Labs experimentally verified Read's principle when they achieved a pulsed power output of 80 mW at 12 GHz from a silicon junction diode driven into avalanche.[6] Advances since 1965 have been so rapid that today avalanche-diode oscillators are established as one

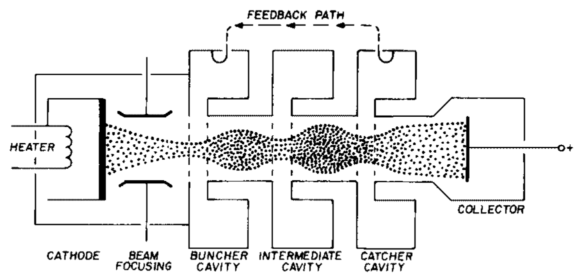



**fig. 3. Klystron operation. Electrons are emitted by the hot cathode surface, formed into a beam, and drawn through the buncher cavity toward the positive anode. As the electron beam pours through the buncher grid it induces an rf voltage in the resonant buncher cavity which, for one half cycle, is in a direction that tends to speed up the electrons flowing through the gap; on the following half cycle the electric field tends to slow down the electrons as they cross the gap. This is called "velocity modulation." In the drift tube the electrons which have been speeded up tend to overtake the electrons which have been slowed down during the preceding half cycle, forming clumps or "bunches" of electrons in the drift tube. If the velocity of the electron beam and the length of the drift tube are properly adjusted, the bunched electrons will be completely formed by the time they reach the gap of the catcher cavity. The time between arrival of individual electron bunches coincides with one rf cycle. When a feedback path is provided, the klystron becomes a self-excited oscillator. Intermediate cavities serve to remodulate the electron beam and greatly increase the gain of the device.**

*Although not discussed in this article, bipolar transistors, GaAs fets, varactor multipliers, and tunnel diodes are widely used on the microwave frequencies. The upper frequency limit for bipolar transistors is now about 4000 MHz, but GaAs fets (called "gas fets") have been used experimentally at frequencies as high as 15 GHz. Varactor multipliers are used in many microwave applications where good frequency stability is required, and tunnel-diode receivers are used commercially on the frequencies from about 4000 MHz to 15 GHz.

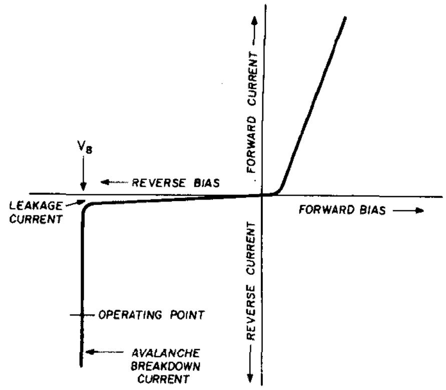



**fig. 4. Voltage-current plot for typical junction diode. When reverse bias exceeds the breakdown voltage, $V_B$, diode is biased into the avalanche breakdown region where large numbers of electrons are generated by secondary emission (also called avalanche multiplication).**

of the most important of the solid-state microwave power sources. Since the operation of the Read device is based on a combination of impact avalanche breakdown and electron transit-time effects, diodes of this general type are generally called IMPATT diodes from IMPact Avalanche and Transit Time.

## IMPATT operation

The basic construction of an IMPATT diode is similar to that of any pn junction diode. Shown in fig. 4 is a typical plot of dc current vs voltage for a pn junction diode. When the diode is forward biased, current increases rapidly for voltages above 0.5 volt or so. When the diode is reversed biased, however, a very small current called "leakage current" flows until the breakdown voltage, $V_B$, is reached. In ordinary rectifier diodes the breakdown voltage, $V_B$, determines the maximum PIV rating of the diode; if this rating is exceeded the diode will be destroyed.

When a pn junction is reverse biased a depletion region forms in the n-type region of the diode with its width depending on the applied bias. If the bias voltage is less than $V_B$ the depletion zone acts like a nonlinear capacitor – this is the property used in varactors and tuning diodes. When the reverse voltage exceeds $V_B$ by a small amount, the diode is biased into the avalanche region. In this region the small leakage current has a very high probability of creating additional electrons by the process of secondary emission or avalanche multiplication. Biasing some diodes into the avalanche region results in catastrophic failure, because once started, the avalanche current cannot be stopped. However, if the semiconductor material is properly doped, the avalanche process can be controlled (as it is in Zener diodes and avalanche rectifiers).

Fig. 5 is a schematic representation of the electron movement in a reverse-biased pn junction with a large number of electrons generated in the avalanche zone flowing into the *drift* zone. In this condition, a large current can flow in the reverse direction with little increase in applied voltage. This is the avalanche breakdown current depicted in fig. 4. If, in addition to the bias voltage, an rf voltage exists across the depletion region of the diode (as it would be if the diode were mounted in a resonant cavity), under certain conditions the rf voltage induces an rf current that is out of phase with the applied voltage. If the rf current in the external circuit lags the rf voltage by more than 90 degrees, this is equivalent to a negative resistance.

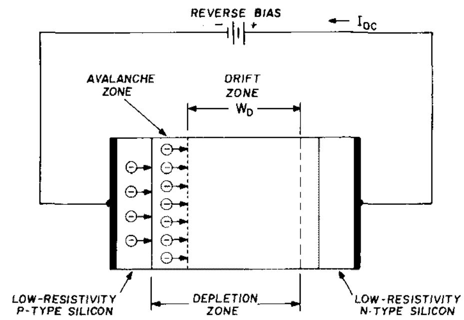



**fig. 5. Schematic representation of a reversed-biased PN junction diode. When biased into the avalanche region, a large number of electrons are launched into the drift region. If the width of the drift zone, $W_D$, is such that the drift time of the electrons is about 37% of the time of one complete rf cycle, the device exhibits negative resistance because the output current will be 180° out of phase with the applied rf voltage (see fig. 6).**

In actual operation the IMPATT diode is biased just above the avalanche point (fig. 4) so the diode is biased into avalanche on positive swings of the rf voltage (fig. 6A). Since the number of electrons generated in the avalanche zone depends not only on the applied voltage but also on the number of charge carriers that are present, the avalanche current pulse continues to build

Low voltage dc to rf
the solid-state way.

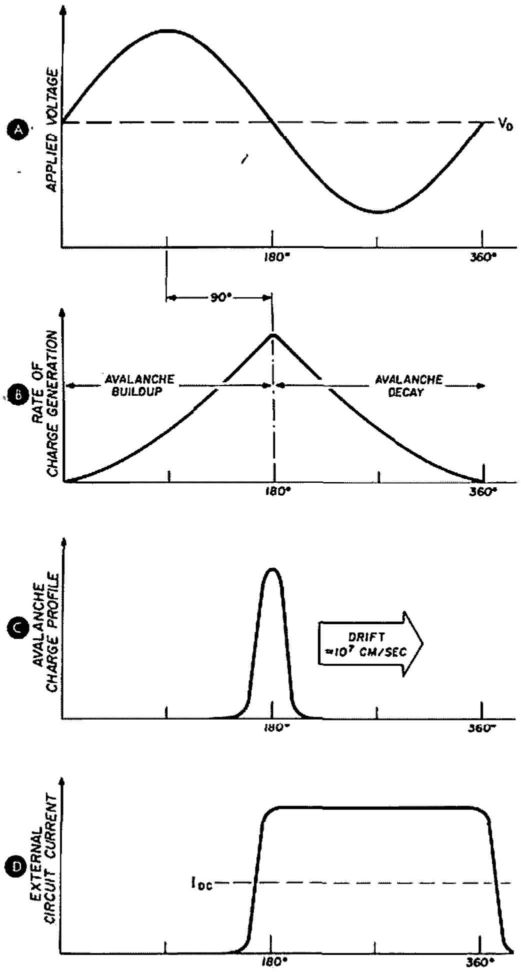



**fig. 6. Voltage and current waveforms in a microwave avalance (IMPATT) diode. When an rf voltage (A) is applied across a reverse-biased PN junction, during the positive half of the rf cycle large numbers of electrons are produced by avalanche generation. Since the avalanche process builds up slowly (B), it peaks when the rf voltage is zero, then slowly decays. This produces a pulse of electrons (C) which drifts toward the anode. The combination of avalanche buildup and drift time produces an external circuit current which is 180 degrees out of phase with the rf voltage (D).**

up even after the rf voltage has begun to drop (fig. 6B). This is because of the highly nonlinear nature of the avalanche generation process. In a properly designed IMPATT diode the excess charge slowly builds up in the avalanche region during the positive half cycle of the rf voltage, and reaches a sharply-peaked maximum in the middle of the rf voltage cycle when the rf voltage is zero. Thus the wave form of the avalanche current, in addition to being very sharply peaked, lags the rf voltage by 90 degrees.

The pulse of avalanche current is launched into the drift zone (fig. 6C) and slowly drifts to the right toward the positively charged n-side of the junction. The electrons drift through the semiconductor material at a nearly constant velocity (about $10^7$ cm/second) so the time it takes them to pass through the drift zone is simply the width of the drift zone, $W_D$, divided by the velocity of the electrons, $v$

$$T = W_{D}/v$$

where $T$ is the drift time. Since drift time is related to the frequency of operation, the width of the drift zone is carefully controlled during the manufacturing process.

While the pulse of electrons is drifting through the diode, they induce an approximate square wave of current in the external circuit as shown in fig. 6D. As can be seen in fig. 7, the combined delay of the avalanche process and the finite transit time across the drift zone has caused a *positive* current to flow in the external circuit while the rf voltage is going through its *negative* half cycle. Therefore the diode is delivering rf energy to the external circuit because of negative resistance.

The useful frequency range of the IMPATT diode is generally above 3000 MHz. Below this point the long transit times require a structure of such thickness that the breakdown voltage is very high. Most high-power IMPATT oscillators are used in the range from 5 to 13 GHz, although they have also been used successfully above 100 GHz. Most of the early IMPATTs were silicon types, but both germanium and gallium-arsenide (GaAs) have been used successfully. All types are noisy because avalanche breakdown is a noisy phenomenon, but GaAs types are somewhat less noisy than silicon devices. This is a problem in some applications (such as receiver local oscillators), but noise can be reduced substantially by proper circuit design. Another disadvantage is the relative high operating voltage (70 to 135 volts), and the requirement for a constant-current power supply. Operating efficiencies are on the order of 12 to 15 per cent, although careful construction, the use of GaAs, and the so-called double-drift structure can increase efficiency to 25 or 30 per cent.

The double-drift IMPATT diode has four layers instead of the usual three because an additional drift region is implanted in the diode (fig. 8). In the single-drift IMPATT described previously, the output current was the result of drifting electrons. The avalanche process, however, generates holes (positive charges) as

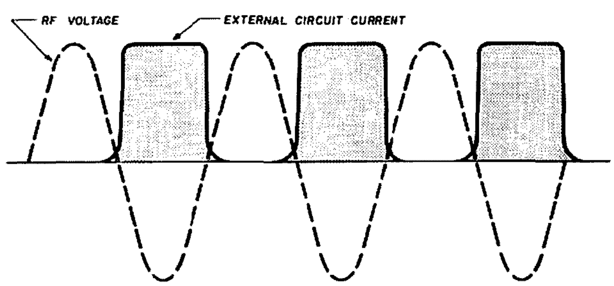



**fig. 7. IMPATT diode exhibits negative resistance because the external circuit current is 180 degrees out of phase with the rf voltage. Because of its negative resistance, it directly converts dc power to *microwave energy*. Mechanism of *microwave* current generation is shown in fig. 6.**

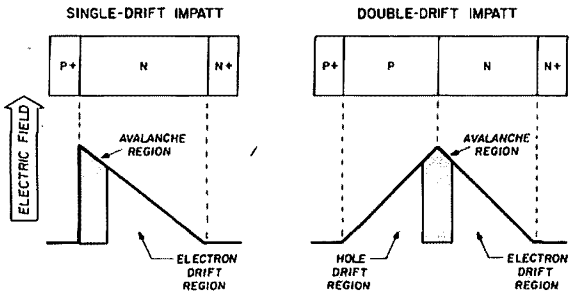
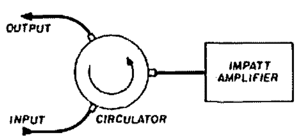



**fig. 8. Construction of single- and double-drift IMPATT diodes. Efficiency is greatly increased in the double-drift diode because the holes (positive charges) generated in the avalanche region drift across the P-doped region in phase with the electrons, providing greater power output. In the single-drift IMPATT the holes are simply returned to the cathode.**

well as electrons. In the single-drift IMPATT the holes simply are returned to the cathode – in the double-drift structure the holes drift across the added p-doped drift region in phase with the electrons, resulting in greater rf power outputs.

Even with operating efficiencies of 25 per cent, heat dissipation becomes the limiting factor when substantial amounts of rf power are required from an IMPATT diode. With good heatsinking techniques the power output can be increased as much as 20 per cent – diamond heatsinks, which have superb thermal characteristics, are being used extensively in commercial IMPATT applications. It should also be noted that IMPATT diodes are being used almost exclusively as amplifiers – few are used as power oscillators.* When used as an amplifier the IMPATT is coupled into the system through a circulator as shown below. The IMPATT amplifier has only one




REFLECTION AMPLIFIER

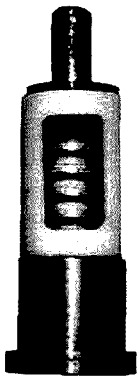
port so the circulator, which allows rf energy to propagate in only one direction, is required to isolate the input signal from the output signal. Operation is similar to that used in a parametric amplifier and is called "reflection amplification."

## TRAPATT diodes

In 1967 researchers at RCA were trying to develop an avalanche diode that would provide operation around 1000 MHz. According to the drift-time theory, as noted above, IMPATT diodes could not be made to operate at that low frequency, but the engineers had hopes of exciting the diode into some other mode. Within a few months they had found a new mode of operation that had both good efficiency and high power output: 425 watts peak, pulsed output with an efficiency of about 25 percent.[7] As researchers continued to work with the new mode, they found a few diodes with efficiencies as high as 60 percent. They also worked on tuned circuits and eventually developed one that permitted continuous tuning from 900 to 1500 MHz. Since the operation of the high-efficiency diode didn't fit any then-known theory, they called it the "anomalous" mode.

Further work at Bell Labs with computer simulation led to the announcement that the high efficiency resulted from the creation of a trapped voltage plasma state between successive sweeps of the classical IMPATT cycle. According to the Bell Theory, this dense plasma then shielded the interior of the diode from the external voltage so the charges (electrons and holes) drift out of the diode at low velocities, causing very long transit times. This led to the acronym TRAPATT for TRApped Plasma Triggered Transit.

**TRAPATT diode configuration built by RCA in 1970 that provided 1200 watts peak power output at 1100 MHz with about 25% efficiency. Five TRAPATT diodes are stacked in a 1N23 rectifier package.**



Quantum physicists from Cornell University didn't agree with the theory offered by Bell Labs – they held that though a trapped plasma could undoubtedly be created in an over-driven diode, it was not fundamental to high efficiency. Their contention was that the parametric theory of Avalanche Resonance Pumping (ARP) had broader validity. Many researchers felt that Bell Labs' TRAPATT and Cornell University's ARP were actually two aspects of the same thing. RCA apparently preferred the TRAPATT model but stayed out of the battle of the acronyms. However, in a moment of humor during one heated debate, someone suggested that the original RCA terminology "anomalous" mode could stand for "A Non-Ohmic Maximum Allowable Large Output Uhf Source!"

In the early 1970s it was discovered that many ordinary silicon rectifier diodes could be made to oscillate in the TRAPATT mode,[8,9] but since this mode of operation is very fussy and requires tricky resonant circuit design to generate the required voltage waveform and suppress higher mode harmonics, the device has found limited applications – primarily in military L-band radar systems.

*Researchers in France have reported that the IMPATT diode operates with very high efficiency in diode frequency-multiplier circuits, but few details are available. This application may be widely used in the future.

## Gunn devices

In 1963, when John Gunn of IBM was studying the bulk resistance of a sample of n-type gallium arsenide, he discovered that when the voltage impressed across the sample was raised above a certain point (**fig. 9**), the current became unstable and began to pulsate cyclically at microwave frequencies.[10] The mechanism which caused this to happen was a mystery at first, but Gunn suspected that a negative resistance was probably responsible, and suggested that a decrease in the mobility (velocity) of the electrons with an increase in applied voltage could account for the negative resistance. This was eventually proved to be the case.

Unlike most other materials, the electrons in gallium arsenide (GaAs) can be in one of two conduction bands, one with much higher electron velocity than the other. As the voltage across the GaAs in increased, more and more electrons are scattered to the low mobility band. This is shown graphically in **fig. 10**. Below the threshold point the current through the material is proportional to the applied voltage, so it behaves as a resistor. As the voltage is increased above threshold, however, sufficient electrons are displaced from the high mobility band that the net electron velocity through the GaAs begins to drop. Since the current through the material is proportional to electron velocity, this means a GaAs resistor (Gunn diode) will have a region of

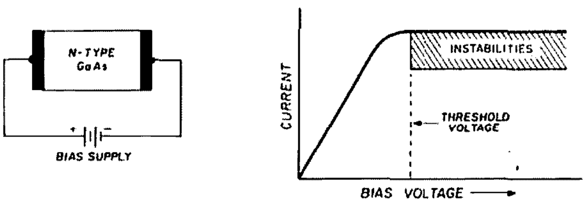
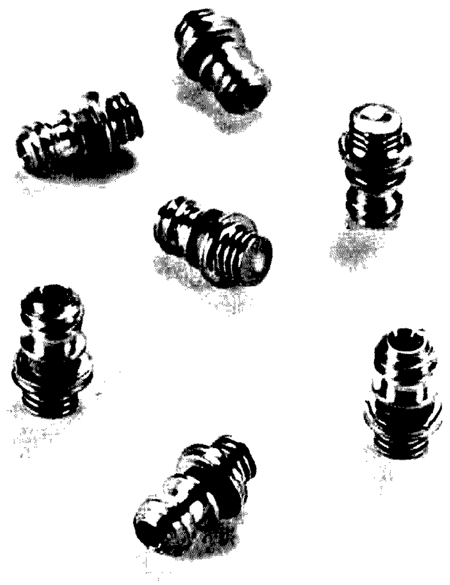



**fig. 9. When the bias voltage across a sample of n-type gallium arsenide (GaAs) is increased above the threshold point, the current becomes unstable and pulsates in a cyclic way. This is a result of negative resistance and is called the Gunn effect.**

negative resistance — decreasing current with increasing voltage. As the voltage is increased past the negative resistance region, the current flow once again increases with applied voltage. Since this behavior is based on the transfer of electrons from one conduction band to another, microwave oscillators of this type are often called Transferred-Electron Oscillators or TEOs.

Although the negative resistance of GaAs accounts for its current instability characteristics at certain bias levels, the oscillation at microwave frequencies requires further explanation. As was mentioned previously, when the applied voltage is below threshold, the GaAs behaves as a linear positive resistance; under these conditions the internal electric field is constant throughout the material as shown in fig. **11A**. If the applied voltage is increased above threshold, many of the electrons entering at the cathode are entering faster than they leave. This leads to a high field buildup at the cathode with an accumulation of electrons on the cathode side and a depletion of



**Construction of Gunn diodes. Most of the package shown here is actually the neatsink — the ceramic-packaged device is only 0.025" (0.6mm) thick. To give you an idea of size, the stud is threaded 3-48 (M2.5).**

electrons on the anode side (fig. **11B**). The electric field throughout the rest of the material begins to fall to a value below threshold. The high field domain drifts rapidly across the GaAs wafer (figs. **11C** and **11D**) to the anode where it is collected (fig. **11E**). When the domain reaches the anode the bias supply again causes the field at the cathode to exceed the threshold level — a new domain is formed and the process repeats itself.

The current through the GaAs is lower during the transit time of the domain, and increases momentarily

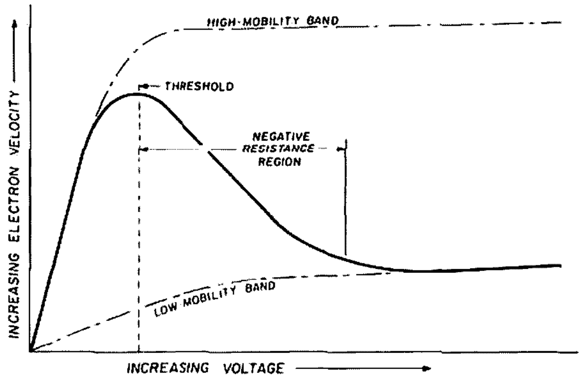



**fig. 10. Electron velocity in GaAs as a function of applied electric field. As the electric field is increased, more and more electrons are scattered to the low mobility band, resulting in a net decrease in current flow through the material — this is equivalent to a negative resistance.**

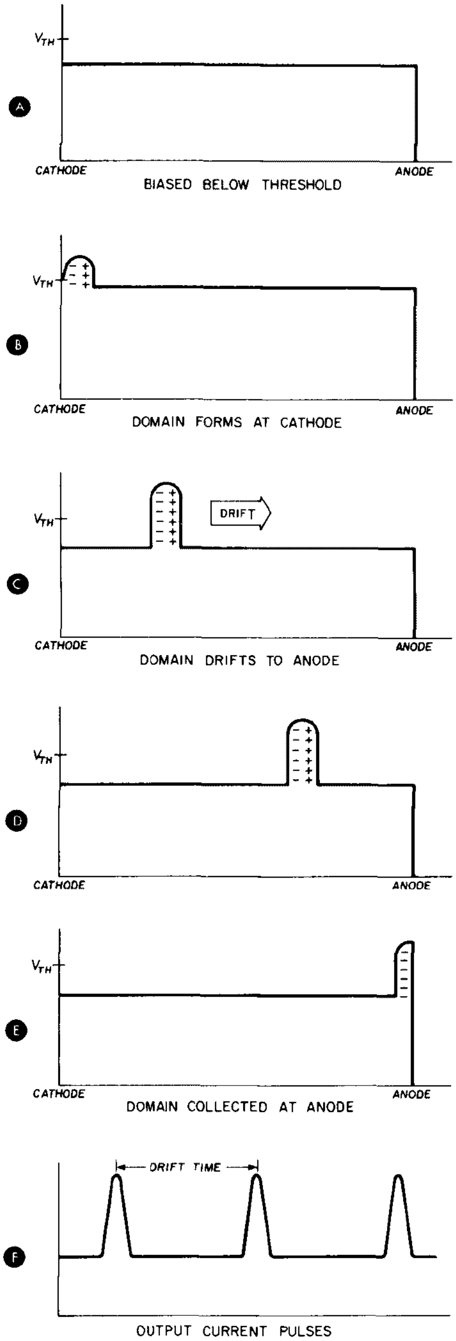



**fig. 11. When a wafer of GaAs is biased below threshold, $V_{th}$, the material behaves as a linear resistance (A). A charge layer (domain) forms at the cathode when the material is biased above threshold (B), and drifts toward the anode at about $10^7$ cm per second (C) and (D). Note that when the field domain forms at the cathode, the electric field in the rest of the material drops below the threshold level. When the domain is collected at the anode (E), the field in the material momentarily increases above threshold, a new domain forms at the cathode, and the process repeats itself. The current through the GaAs is lower during the transit time of the domain and increases when the domain reaches the anode, giving a series of sharp current spikes (F).**

when the domain is extinguished at the anode. Thus the output is a series of narrow current spikes with a period equal to the transit time through the wafer (fig. 11F). The domain velocity is about $10^7$ cm per second so the wafer of GaAs must be about 10 microns thick for operation at 10 GHz. Since the frequency of the output current pulses is a function of drift time, this is called the transit-time mode of operation. It is rarely used, however, because frequency tuning is nil and efficiency is very low.

In practical microwave circuits the Gunn device is mounted in a high-Q resonant circuit – this provides a considerable tuning range because the rf voltage in the circuit influences the formation of the field domain at the cathode by swinging the applied voltage above and below the threshold level as shown in fig. 12. If the bias is set so the rf voltage drops below the sustaining voltage, $V_s$, the field domain will be quenched; when the rf voltage rises above threshold, $V_{th}$, a new domain will form at the cathode. As the domain starts to drift across the GaAs, the rf voltage will increase to a maximum and then decrease until it, too, is quenched.

The delayed-domain mode occurs when the instantaneous rf voltage never falls below the sustaining value but does go below threshold for a part of the cycle. If the field domain reaches the anode when the applied voltage is between $V_s$ and $V_{th}$, the formation of a new domain is delayed until the rf swings above threshold. Using these techniques and others, the frequency of the output current spikes will adapt to the resonant frequency of the external tuned circuit and can be tuned over very wide frequency ranges.

The greatest advantage a Gunn device has over IMPATT and TRAPATT diodes is its ability to operate over a wide band with less noise and lower bias voltages at equivalent frequencies – this is an important consideration for amateurs who want to build simple microwave systems that operate from batteries for portable,

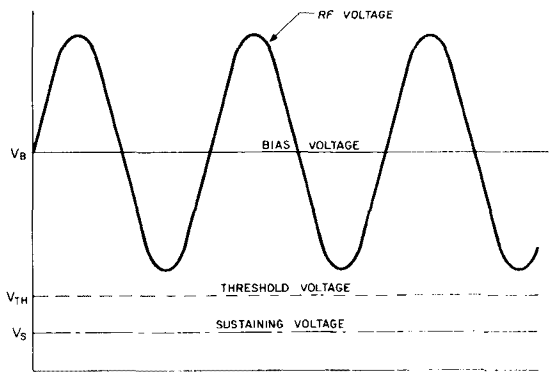



**fig. 12. Operating mode of a Gunn diode in a high-Q resonant circuit is determined by the bias level, as shown here. If the diode is biased so the rf voltage drops below the sustaining voltage, $V_s$, the field domain will be quenched until the rf voltage rises above threshold, $V_{th}$, and the formation of a new domain at the cathode will be delayed (see text). Therefore, the frequency of the output current spikes will adapt to the resonant frequency of the tuned circuit and can be tuned over wide frequency ranges.**

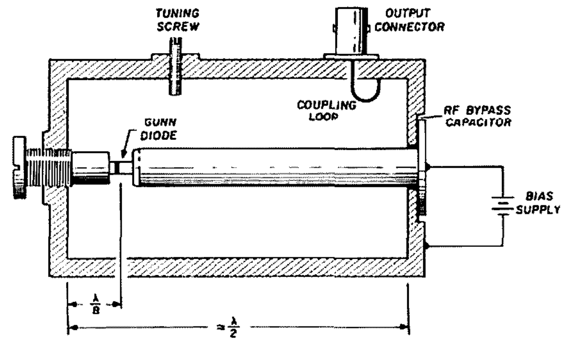



fig. 13. Simple Gunn-diode oscillator uses a half-wavelength coaxial cavity. Impedance matching is provided by the output coupling loop. This type of circuit can be tuned over an octave or more, but difficulties with oscillation at harmonic frequencies are common, and the coaxial cavity is more sensitive to temperature changes and load mismatches than waveguide resonators.

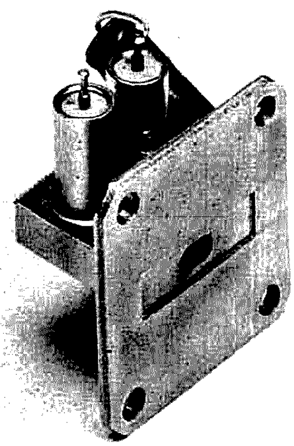
nology, fresh from the laboratory; the GaAs manufacturing process was still in its infancy so there were problems with the diodes, and nobody had any comparable experience to fall back on. Gunn diodes have improved over the years, and we now know what types of resonant circuits work best, but the plain truth of the matter is that not much actual circuit design takes place — not in the traditional sense anyway. As one designer at Microwave Associates pointed out recently, "Those who have experience with Gunn diodes usually take a rough cut at what they think will work, and then home in on the final layout by trial, error, and a good deal of sheer feel."

One of the simplest forms of Gunn oscillator circuits is shown in fig. 13. Here, the diode is mounted at one end of a half-wavelength coaxial cavity which is adjusted

mountain-top expeditions. Moreover, Gunn diodes are easily frequency modulated and lend themselves to automatic frequency control (afc). Compared to tubes the Gunn device operates at lower temperatures and without a vacuum, factors that contribute to longer life. In fact, a number of manufacturers are predicting useful lives of well over 300,000 hours for CW devices (in case you don't want to figure it out, that's equivalent to about 34 years of 24-hour-per-day operation). On the debit side, the Gunn diode is less efficient and has lower power output then other solid-state microwave devices, but this is more than offset by its simplicity of operation, wide tuning range, and lower operating voltage.

## Gunn oscillators

Many early attempts to build Gunn oscillators were not all that successful — the results were seldom reproducible. Sometimes the device refused to oscillate in the resonant circuit, or if it did oscillate it wouldn't be on the desired frequency, but this was new tech-



Microwave Associates varactor-tuned Gunn oscillator for tne amateur 10-GHz band. Terminals are for the Gunn power supply and varactor bias. Power output is 20 milliwatts.

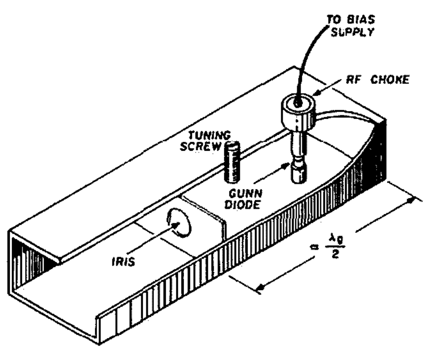



fig. 14. Simple waveguide resonator for Gunn-diode oscillators. In this circuit the microwave energy from the Gunn diode is coupled into the cavity with a post mounted between the narrow dimension of the waveguide. The size of the opening (iris) is optimized for maximum power output and isolation from impedance mismatches. The rf choke requires careful design for minimum rf loss.

to the desired operating frequency with a tuning screw. The location of the output coupling loop determines the load impedance presented to the diode. This type of resonator is easy to build and can be tuned over a very wide frequency range, typically an octave, but it has several disadvantages. For one thing, the Q is relatively low and the diode may want to oscillate at a harmonic frequency. In comparison with waveguide cavities, the coaxial resonator is also more sensitive to temperature changes and load mismatches.

For most applications a much better choice is the post-coupled rectangular waveguide cavity shown in fig. 14 separated from the output waveguide by a coupling iris. The size of the iris is determined experimentally for the best compromise between maximum power output

table 1. Operating characteristics of the Microwave Associates MA-87127 series 10-GHz Gunnplexers.

| | |
|---|---|
| RF center frequency | 10250 MHz (4V varactor bias) |
| RF power output | 20 mW |
| Tuning: mechanical | ± 100 MHz |
| electronic | 60 MHz minimum |
| Frequency stability | −350 kHz/°C maximum |
| RF power vs temperature and tuning voltage | 6 dB maximum |
| Frequency pushing | 15 MHz per volt, maximum |
| Noise figure | 12 dB maximum |
| Input requirements: | |
| Gunn voltage | +10 Vdc typical |
| Gunn current | 500 mA maximum |
| Tuning voltage | +1 to +20 volts |
| Rf connectors | mates with UG-39/U waveguide |
| Operating temperature | −30 to +70 °C |

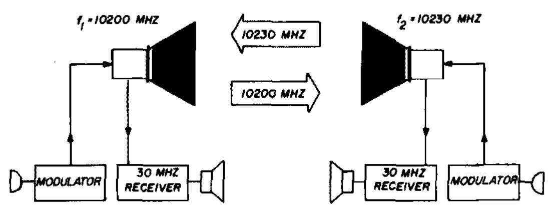



fig. 16. Gunnplexer operation. Since the same oscillator is used as both a transmitter and local oscillator for the mixer, the i-f at eacn end of the link must be at the same frequency, and the frequencies of the Gunn oscillators must be separated by the i-f. In the example shown here the Gunnplexer at one end of the link is tuned to 10200 MHz — 30-MHz i-f receivers are used so the Gunnplexer at the other end must be tuned 30 MHz higher or lower (10230 or 10170 MHz).

and isolation from changes in diode impedance and load. The tuning rod may be either metal or a low-loss dielectric. The diode must be properly decoupled from the bias supply to minimize any rf resonances in the bias circuit and to prevent any rf loss. None of these things is trivial, so if you're interested in building your own Gunn oscillators for the 10-GHz amateur band, I suggest that you try one of the proven designs published in the RSGB's *VHF - UHF Manual*.[11]

If you don't have any previous experience with microwave circuits, you may find it easier and less frustrating to purchase one of the new 10-GHz *Gunnplexers* which are being offered to amateurs by Microwave Associates.* These Gunn-oscillator transceivers provide 20 mW of output power, include a built-in low-noise Schottky mixer diode, and are provided with varactor tuning. A ferrite circulator isolates the receiver and transmitter functions. The electrical specifications of the Gunnplexer are listed in table 1; a cut-away view of the transceiver is shown in fig. 15.

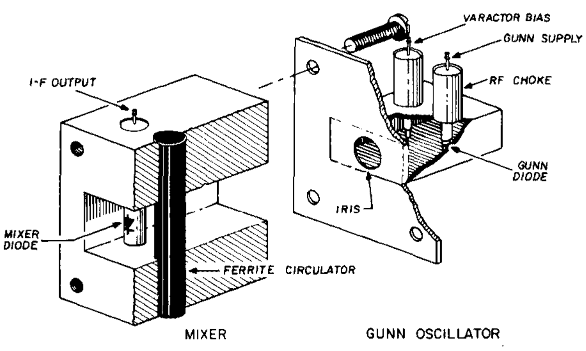



fig. 15. Cutaway view of the 10-GHz Microwave Associates Gunnplexer. The post-coupled Gunn diode is tuned to the desired frequency with the dielectric tuning screw, and the rf energy is coupled out through an iris. The ferrite circulator couples a small amount of energy into the Schottky mixer diode and isolates the transmit and receive functions. Mixer injection can be adjusted with the small screw mounted in front of the circulator. A horn antenna provides 17 dB gain.

In the Gunnplexer the Gunn oscillator provides both the transmit power and LO injection for the mixer diode. Therefore, the i-f used at each end of a communications link must be tuned to the same frequency, and the frequencies of the Gunn oscillators at each end of the link must be separated by the intermediate frequency. This is the same system used with klystron polaplexers and is shown in fig. 16. If the i-f is at 30 MHz, for example, and one Gunn oscillator is tuned to 10200 MHz, the other oscillator is tuned 30 MHz higher (or lower) to 10230 MHz (or 10170 MHz). Gunnplexers can also be used for two-way communications with stations which use polaplexers or separate 10-GHz transmitters and receivers. If a polaplexer is used at one end of the link, however, you should expect about 3 dB loss because of the difference in polarization.

All you need to put the Gunnplexer on the air is a well regulated 9-volt power supply (200 mA maximum), a bias supply for the varactor, an fm receiver, and a microphone and speech amplifier. This is the system used by W1HR and W1SL in their first two-way communications with Gunnplexers. Later you may want to add automatic frequency control (afc) or a phase-locking system, but this isn't necessary to get started.

The 30-MHz i-f has been the standard for amateur microwave communications for a number of years, but there's no reason why you can't use a standard 88-108 MHz fm broadcast receiver as a tunable i-f. If you choose to go this route, be sure to pick a frequency that's not occupied by a nearby fm transmitter — otherwise you may have problems with i-f feedthrough. In metropolitan areas it may be impossible to find a clear

*The Microwave Associates MA87127 *Gunnplexer* is a complete 10-GHz transceiver consisting of a Gunn oscillator, tuning diode, detector, and circulator and is priced at $85. Also available is the MA87108 which consists of the Gunn oscillator and tuning diode ($60), and the MA87140 which includes a complete transceiver and a 17-dB gain horn antenna ($108). Two complete transceivers with horn antennas, part number MA87141, are priced at $185. Write to Microwave Associates, Inc., Burlington, Massachusetts 01803 for the name of your nearest sales representative.

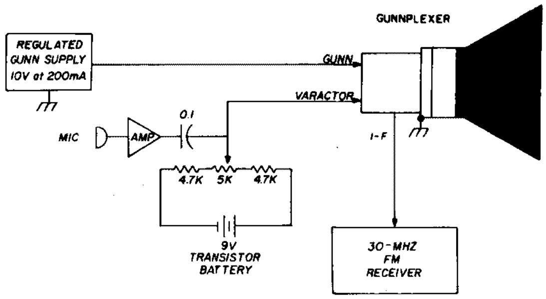



**fig. 17. Basic system used by W1HR and W1SL for short-range communications with the 10-GHz Gunnplexers. Output frequency is adjusted ±30 MHz with the 5k potentiometer across the 9V varactor bias supply.**

channel, but most fm receivers can be "tweaked" slightly so they will tune to a clear spot above or below the fm broadcast band. If you use an fm broadcast receiver you won't be able to work stations using a 30-MHz i-f, but this is an inexpensive way to get started, and you can always to to·30-MHz i-f later.

If you decide to use a 30-MHz i-f, military surplus i-f strips are inexpensive, or you can use a tunable fm receiver such as the old Hallicrafters S-36 or S-27, or the military surplus BC683. W1SL and I used 35-year-old S-36 receivers in our initial experiments, but are planning to build some solid-state replacements in the near future. With the wide availability of ICs designed for rf amplification and fm demodulation, this should be a relatively easy task.

## communications range

One of the first questions you're probably asking is what kind of communications range can I expect with a

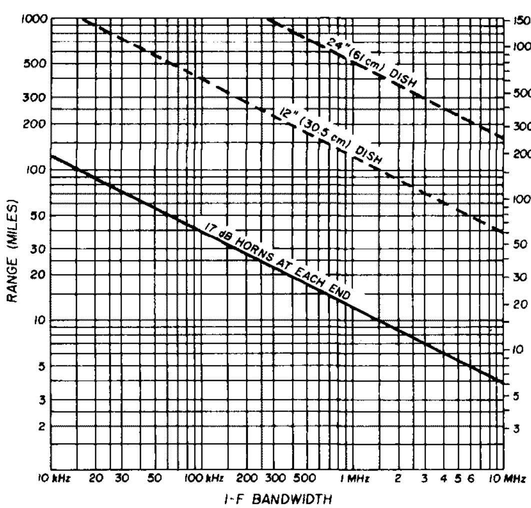



**fig. 18. Range vs i-f bandwidth for the Microwave Associates 10-GHz Gunnplexers. Dashed lines show increased range available with 12-inch (30cm) and 24-inch (61cm) parabolic reflectors.**

20 mW Gunnplexer system? As shown in **fig. 18**, this is a function of the bandwidth of the i-f system. This graph assumes a noise figure of 12 dB at 10 GHz which should be no problem with a low-noise (about 1 dB) i-f strip, and 17-dB horn antennas at each end of the link. For the bandwidth of an fm broadcast receiver, 240 kHz, the *maximum* line-of-sight range is about 25 miles (40 km). However, this graph is based on "threshold" (the beginning of reception of intelligible speech) and allows no margin for fading due to rainfall, multipath propagation, or other environmental effects. Therefore, for a practical system with good signal-to-noise ratios, somewhat less range should be expected. Range can be increased considerably by using parabolic reflectors, as shown by the dashed lines in **fig. 18**, but this entails additional cost and you may have trouble getting the antennas properly lined up.

Range can also be improved by using a narrower i-f bandwidth, but this requires the use of automatic

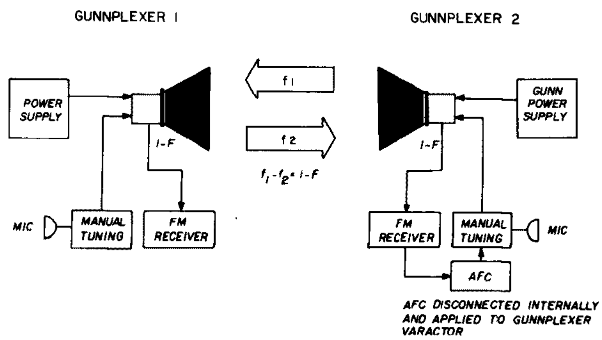



**fig. 19. Basic arrangement for automatic frequency control (afc) of the Gunnplexer. Manual frequency control is used at one end of the link; afc at the other end of the link is derived from the fm receiver and applied to the varactor.**

frequency control or a phase-lock arrangement. Since the drift characteristics of the Microwave Associates Gunnplexer is about −350 kHz per °C maximum (downward drift with increasing temperature), it's doubtful that the 240-kHz bandwidth of an fm broadcast receiver would be practical without continually adjusting varactor bias or using afc. The basic afc system is shown in fig. 19. To prevent the two Gunnplexers from chasing each other all over the band, only one end should use afc of the Gunn diode. Afc of the i-f LO is permissible at both ends of the link.

Stability can also be improved by placing the Gunnplexer in an insulated box and heating it with a lightbulb or other heat source (such as a power resistor). For maximum temperature stability, a proportional temperature control system is suggested (reference 12 describes a proportional control circuit for crystal ovens that could be easily adapted to the Gunnplexer).

For an idea of what can be accomplished with low power on 10 GHz, consider the new world's record on this band which was set in August, 1976, by GM3OXX in Scotland, and G4BRS, in Cornwall, England — a

distance of 324 miles (521 km). Both stations used 10 mW Gunn oscillators with parabolic reflectors – a 24 inch (61cm) dish in Scotland and a 30-inch (76cm) reflector in England. This was their ninth try over this particular path, so it wasn't as easy as it sounds, but it should give you an idea of what can be done with simple equipment, patience, and a *lot of persistence.*

## phase locking

One of the best ways to improve communications range with the Gunnplexer is to decrease i-f bandwidth but this can only be done by phase locking the transmitter to a stable crystal oscillator. I don't have any practical, tried and true circuits to offer yet, because I don't know of any amateurs who have built a phase-locked Gunnplexer system, so the following are merely suggestions.

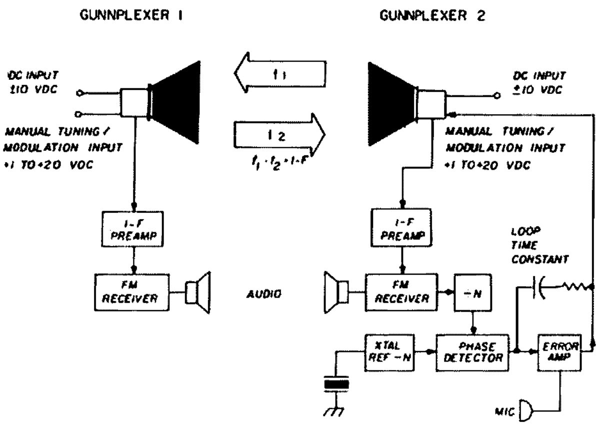



**fig. 20. Gunnplexer phase-lock system suggested by W1FC. Output of i-f amplifier is divided down and compared against a crystal-controlled reference oscillator. Output of the phase detector is amplified to a suitable level to keep the Gunn oscillator locked to the crystal. Time constant of the RC network shunting the error amplifier is chosen to allow the Gunnplexer to be frequency modulated.**

One arrangement, which was suggested by W1FC of Microwave Associates, is shown in fig. 20. In this system the output of the i-f amplifier is divided digitally and phase compared against a stable crystal reference. The dc output information from the MC4044 phase detector is amplified and fed to the Gunnplexer tuning varactor. The time constant of the RC network which shunts the dc error amplifier permits the Gunnplexer to be frequency modulated. Without the RC circuit, the phase-lock system would tend to cancel the modulation.

Another, more complex phase-lock circuit, based on the San Bernadino Microwave Society's *Rocloc* system for klystron polaplexers,[13] is shown in **fig. 21**. In this circuit a small portion of rf from the Gunnplexer is mixed with the output from a harmonic multiplier – the output from the mixer is then phase compared with a low-frequency crystal-controlled source. The rest of the system is similar to that shown in fig. **20**.

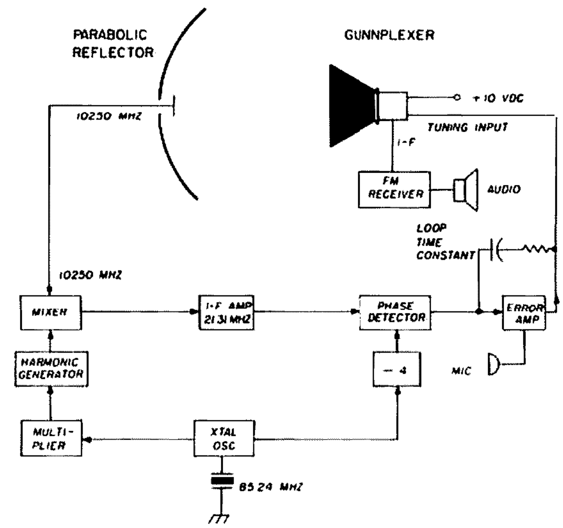
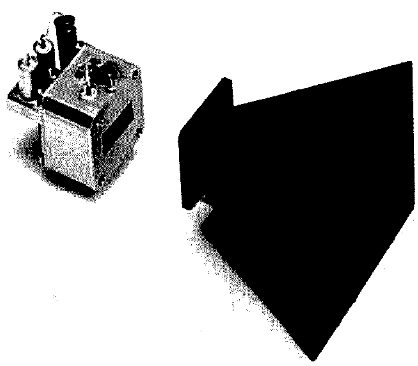
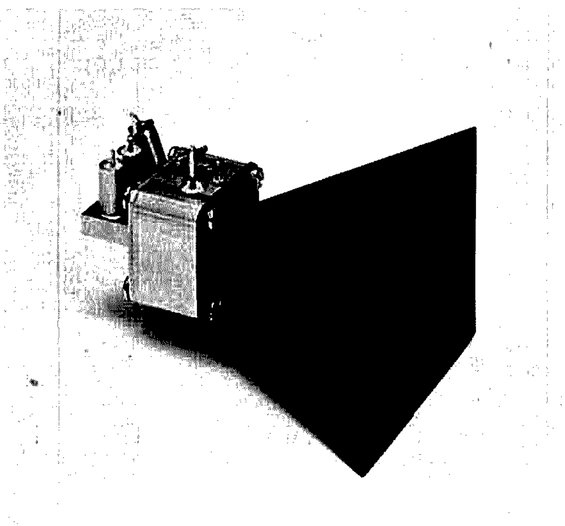



**fig. 21. Proposed phase-lock system for a Gunnplexer operating on 10250 MHz. In this system the 120th harmonic of the 85.24-MHz crystal oscillator is mixed with a sample of rf energy from the Gunnplexer. The difference output is amplified, then phase compared with a signal derived from the crystal oscillator. The error voltage is amplified and fed to the tuning varactor in the Gunnplexer. With proper design, frequency stability of the Gunnplexer will be the same as the crystal reference oscillator.**

Phase locking can also be used to good advantage in the receiver as a tracking filter. Whereas the threshold of intelligible speech in a conventional fm system occurs when the fm signal is about 10 dB above the noise, a phase-locked tracking filter can linearly demodulate fm signals buried 20 to 30 dB in the noise.[14] One IC on the market which was designed specifically for this task is the Exar XR-215.



**Microwave Associates 10-GHz Gunnplexer and 17-dB horn antenna. Receiver section is housed in waveguide section machined from large block of metal. This improves thermal stability of the unit.**



Microwave Associates 10-GHz Gunnplexer.

## i-f circuits

Good afc capture range requires a wider bandwidth than optimum signal processing, so a dual i-f system is recommended. One suggestion here is an input at 30 MHz followed by conversion to 10.7 MHz for signal processing. A parallel i-f at 30 MHz would be used for afc control. The 30-MHz preamp should have a noise figure on the order of 1 dB and should be impedance matched to the Schottky mixer in the Gunnplexer (about 200 ohms). The capacitance of the mixer diode (approximately 27 pF) can be resonated with an rf choke connected from the i-f output terminal to ground (use 1 $\mu$H at 30 MHz, 0.33 $\mu$H at 100 MHz).

## radiation hazard

Although 20 mW isn't very much power, remember that it's concentrated at the small, open end of the waveguide so power density is about 6.2 mW per square cm. This is considerably higher than OSHA's 1 mW/cm$^2$ safety limit. Fortunately, rf power density falls off to safe levels a few inches (15 cm) away, but remember that your eyes are especially susceptible to damage from rf radiation — *never* look into the open end of a Gunnplexer when it's operating.

## summary

The amateur microwave bands have been badly neglected by most amateurs, but this is one area where amateurs can still make contributions to modern rf technology. Presented here are the basic requirements for a modern, solid-state, amateur microwave system, and some suggestions for putting your own Gunnplexer on the air. In future months we will try to publish practical, successful circuits — if you have suggestions for improvements, or have solved a problem that may give others trouble, we would like to have the opportunity to publish it in *ham radio*.

I would like to thank Dana Atchley, W1CF, of Microwave Associates for making Gunnplexers available to the amateur community, and Fred Collins, W1FC (ex W1FRR), and Dr. Ron Posner (ex K6DJB) for their circuit suggestions.

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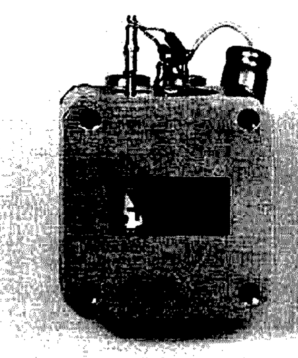
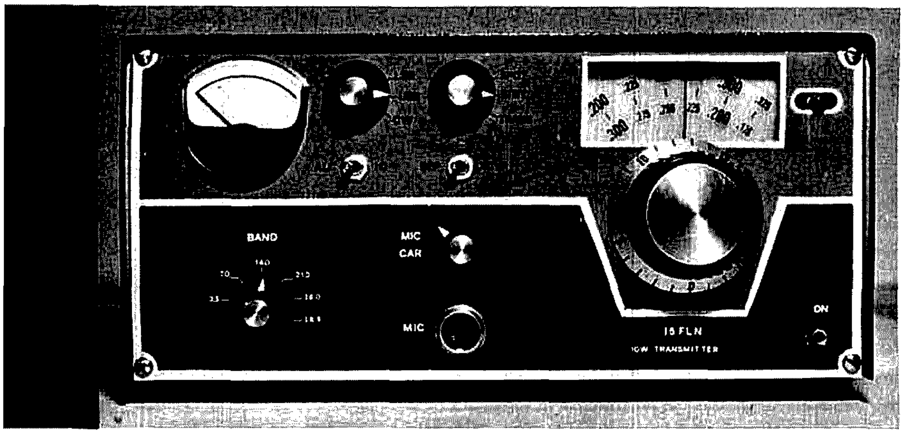


Head-on view of the Gunnplexer showing the mixer diode, left, and ferrite circulator (black cylinder to right). The small screw which protrudes through the top of the waveguide is used to adjust mixer injection.




# solid-state five-band transmitter

## A companion transmitter to the I5TDJ receiver features all bands, solid-state circuitry, and 10-watt output

***On the air, amateurs*** **speak** proudly of using the "S-Line" or the "Drake Line." To complement the receiver I described last year in these pages,[1] I built a companion five-band solid-state transmitter. Thus I felt I could proudly speak of using the "TDJ Line!"

I actually started the design in 1973 when high power rf transistors were not available in Italy. Thus, the output is only a modest 10 watts. None the less, I've made contacts with all continents and had a nice QSO on CW and ssb with JD1ACH during the Ogasawara (Bonin Island) DXpedition.

The transistors I used are a mixture of Phillips and U.S. types, collected over the years. Despite that, the 3rd-order IMD measures −28 dB below one tone of a 2-tone test. Output is actually greater than ten watts on all bands, with no tuning. A SPOT switch and carrier insertion for linear amplifier tuneup are included for convenience. Again, my friend I5FLN has duplicated my circuitry.

### circuit description

The transmitter was designed to transceive with my receiver; the rf mixing scheme being identical. In the block diagram, **fig. 1**, note that the two modules, vfo and vfo converter, are common to both units. The builder has three options: you may use these two modules in the receiver, you can duplicate just the vfo module, or you can duplicate both modules and have a completely self-contained transmitter with frequency control from either unit – just like Drake. Don't forget a light to indicate which dial is operative!

In the ssb generator module, shown in fig. 1, the microphone signal is amplified in two stages and then applied to a diode-ring balanced modulator. The carrier is initially generated at 454 or 455 kHz, for CW/ssb. Undesired products are attenuated and suppressed by adjustment of the modulator and also FL1, a 2.4-kHz mechanical filter. The output from the filter is applied through a buffer to a balanced mixer comprised of two fets. Another pair of crystals at 8545 and 9455 kHz provide the signal, which is simultaneously applied to the fet mixer, for USB or LSB selection. The mixer output. after a three-section LC filter, is now a 9-MHz ssb signal.

In the next module, the 9-MHz signal is buffered and applied to a second diode ring mixer for addition to the signal from the vfo converter. Five wide-band LC filters at the output of the diode mixer select the desired band. The combination of diode mixer and LC filters again make for excellent suppression of undesired mixer products. The signal is now fed to a wide band amplifier and then to the power amplifier (PA) module.

Two transistors are used for the driver in the power amplifier module. The final transistor operates as a linear amplifier and has been protected against thermal runaway. No mismatch protection was deemed necessary. On several occasions the antenna was not connected, and no damage has resulted to date! In order to make the

By Piero Moroni, I5TDJ, Cosseria 10, Florence, Italy 50129 (translated by Joe Darmento, W4SXK)

most of the modest power available, a tuned circuit is used to match the final transistor to a 50-ohm output. Each circuit need only be set once during the initial adjustments. A sample of the output is rectified and used for alc. The 0-1 mA meter reads the supply voltage or may be switched to read alc voltage and thus monitor the output.

should be selected for equal forward resistance at 1 mA of current. Crystals Y1 and Y2 provide the 454/455-kHz carrier. On CW, the 454-kHz crystal causes the carrier to be transmitted 1 kHz above the vfo frequency. No problem — your contact receives a 1-kHz signal in the usb position of his receiver. R1 and C1 are used to null the carrier during the initial adjustment.

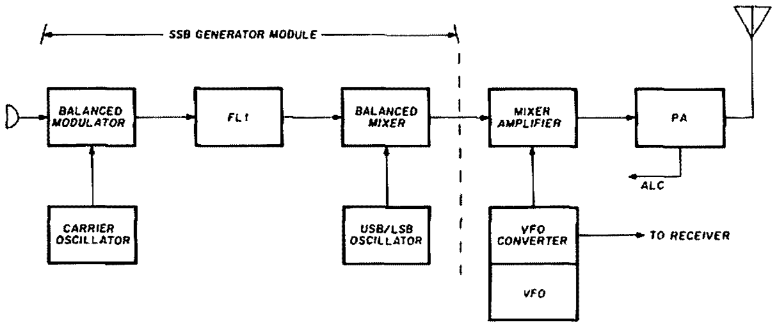



**fig. 1. Block diagram of the solid-state transmitter. The vfo and mixing scheme are identical to the one used in the receiver described by the author in the October, 1975, issue of ham radio.**

Except for the auxiliary circuit, each module was enclosed in a standard aluminum box, 3 x 4¼ x 2 inches (77x107x49mm). The PA module is slightly larger, 3 x 5.7 x 2 inches (77x145x49mm). Each module may be arranged as desired to fit a commercial case, as done by I5FLN. All circuitry is mounted on single-sided epoxy-glass PC boards. I used an isolated pad drill rather than printed-circuit technique; it's easier and quicker for a one-time project. Signal interconnections are made with phono jacks and plugs. All other interconnections enter the modules through 1500-pF feedthrough capacitors. Points designated as +12 always have power applied; +12T implies voltage on with PTT. When winding the toroids, make sure you twist the wires 7 to 8 turns per inch (2 to 3 turns per cm).

## ssb generator module

Referring to fig. 2, Q1 and Q2 amplify the microphone signal to a few hundred millivolts for application to the balanced modulator. The four 1N270 diodes

**Mixer and amplifier module. The individual band filters are mounted on the vertical circuit board. The balanced mixer is on the left and the output circuits on the right.**

The gain of Q3 is controlled by the alc developed from the final transistor. Q6 and Q7 form the balanced mixer for signal conversion to 9 MHz. R2 is the balance pot, and when properly adjusted the mixer, in conjunction with L1, L2 and L3, will provide good attenuation of spurious products without resorting to an expensive commercial filter. One precaution, do not mount any oscillator components near the ends of the mechanical filter or you will not obtain good carrier suppression. Component placement is shown in the photograph.

## mixer amplifier module

In the schematic, fig. 3, transistor Q1 is a buffer amplifier; the 100-ohm potentiometer adjusts the 9-MHz signal level into the diode mixer. Unfortunately, this type of mixer attenuates the desired signal approximately 6 dB and requires about 5 mW of oscillator injection. However, it does attenuate the oscillator signal at least 30 dB with satisfactory performance from 0.5 to 50 MHz. This latter characteristic is eminently desirable to provide constant output on all bands. Thanks to these characteristics the simple filters on the output are perfectly adequate. Except for component values these five filters are all identical to the 3.5 to 4 MHz filter shown in detail in the schematic. The filters are diode selected by the bandswitch. With feedback to improve linearity, the signal is amplified to about 50 mW by Q2 and Q3.

Components for this module are mounted on three small boards approximately 3/4 x 1 inch (15x25mm). One board holds the buffer and mixer, the second the 5 filters, and the third board contains Q2 and Q3. Separating the boards is helpful in home brewing. In case you make a fatal error, you ruin only a small part of your work.

## power amplifier module

Two transistors in push-pull, fig. 4, raise the signal level to 0.5 watt. CR1 is in thermal contact with one transistor to maintain collector currents within safe

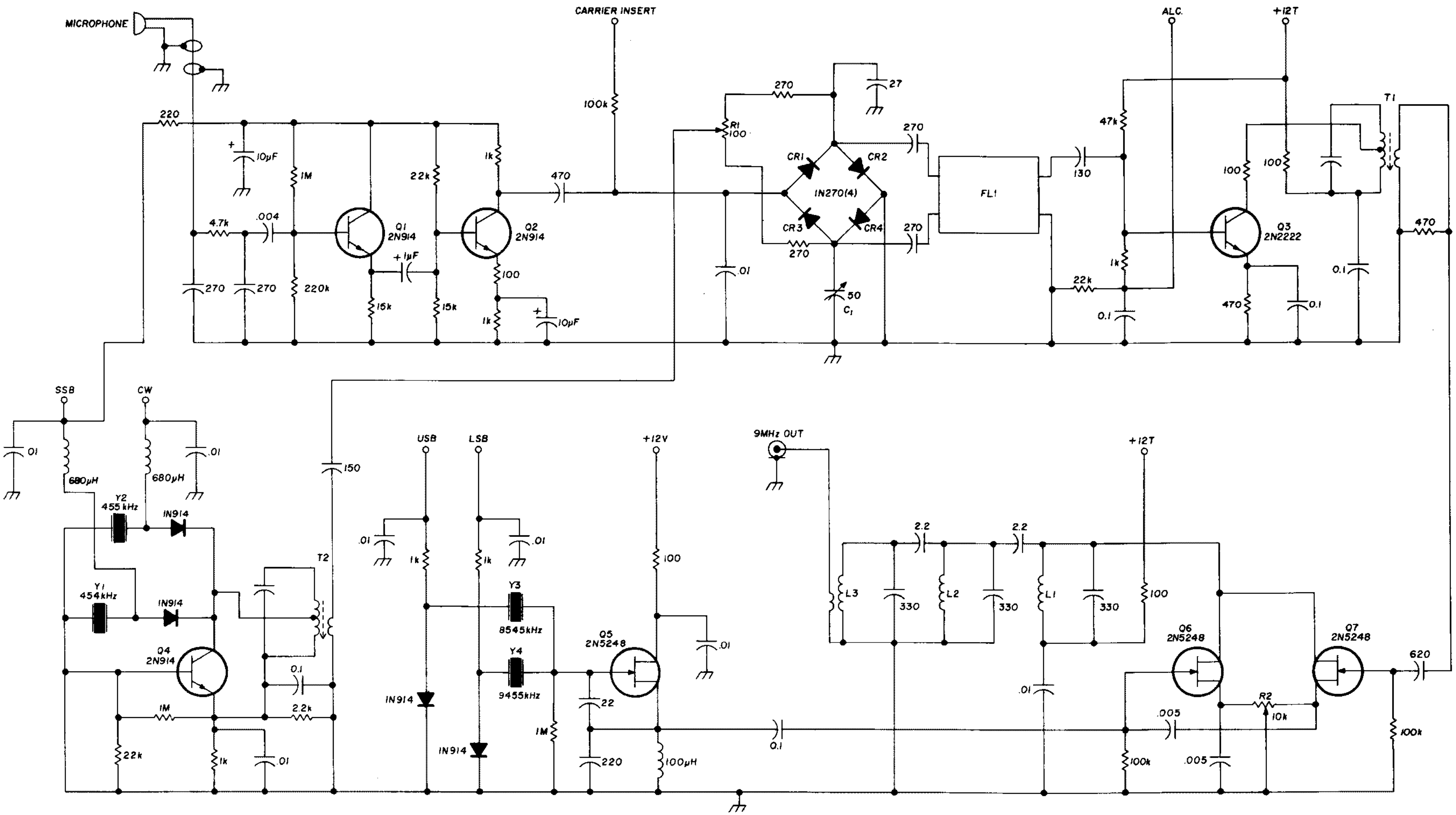



fig. 2. Schematic diagram of the ssb generator. C1 is a 50-pF (ARCO) compression trimmer. FL1 is a Collins F455-Z7 filter. L1 through L3 are each 15 turns of no. 22 AWG (0.6mm) wire wound on a Micrometals T44-6 core. L3 also has a one-turn link. Each coil should be 0.95 $\mu$H. T1 and T2 are 455-kHz transistor i-f transformers (Miller 2031). The crystals are parallel resonant with 32-pF load capacitance.

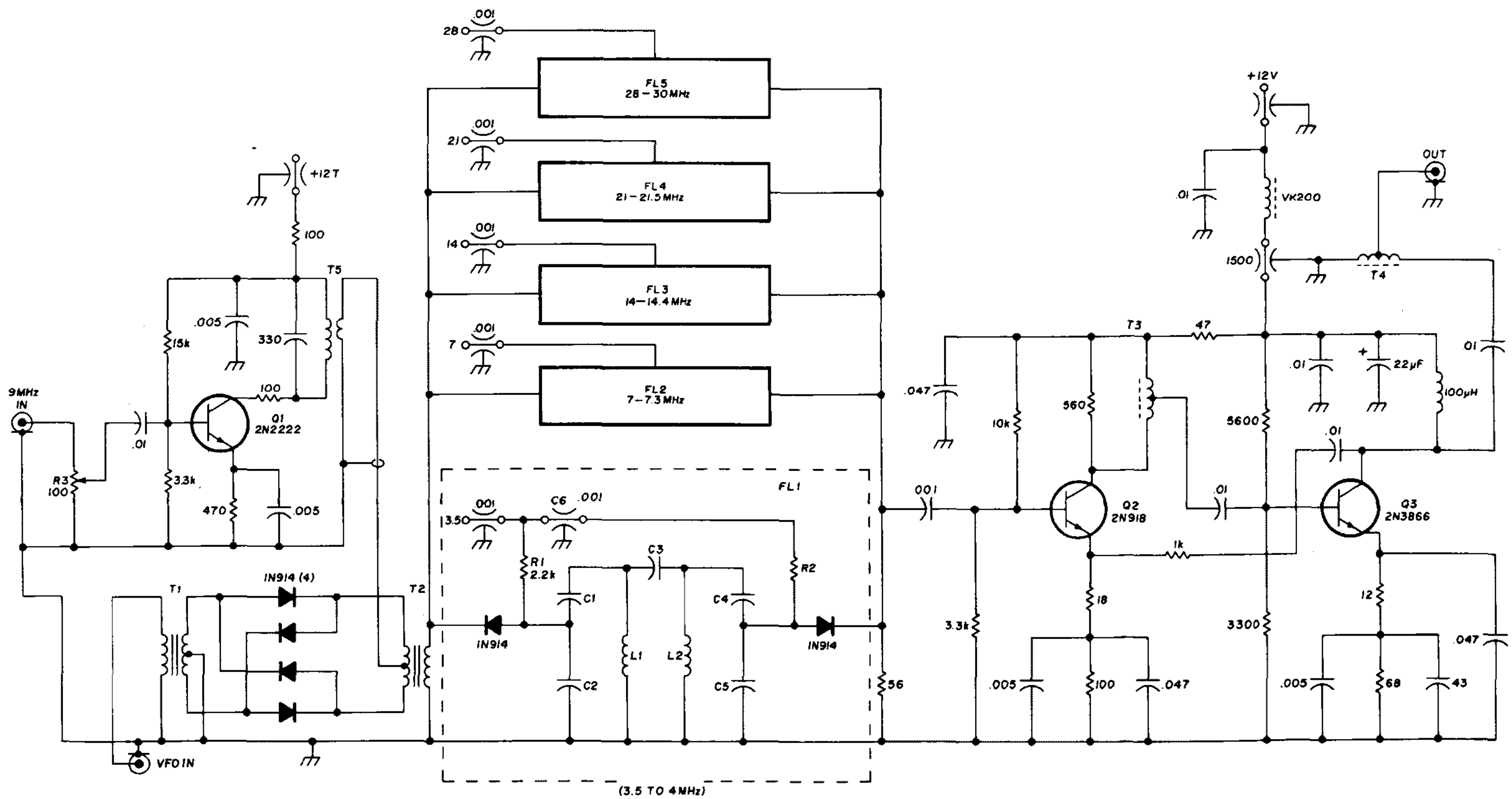



fig. 3. Mixer and amplifier module schematic diagram. T1 and T2 are each 10 turns of no. 32 AWG (0.2mm) trifilar wound on a hi-μ core (Indiana General CF-2). T3 and T4 are identical to T1 except they are bifilar wound. T5 is wound on an Amidon T44-6 core. The primary is 15 turns of no. 24 AWG (0.5mm) with a secondary of 2 turns wound over the ground side of the primary. The rf choke, VK200 19/4B, is available from Elna Ferrite Labs, Woodstock, New York 12498.

| | C1, C4 | C2, C5 | C3 | L1, L2 |
|---|---|---|---|---|
| FL1 | 500 pF | 3000 pF | 39 pF | 3.4 μH, 30 turns no. 24 AWG (0.5mm) |
| FL2 | 330 pF | 3000 pF | 8.2 pF | 1.6 μH, 20 turns no. 24 AWG (0.5mm) |
| FL3 | 150 pF | 1600 pF | 3.3 pF | 0.9 μH, 15 turns no. 22 AWG (0.6mm) |
| FL4 | 100 pF | 1000 pF | 3.3 pF | 0.6 μH, 13 turns no. 22 AWG (0.6mm) |
| FL5 | 68 pF | 430 pF | 3.3 pF | 0.5 μH, 11 turns no. 20 AWG (0.8mm) |

Note: The coils for FL1 to FL3 are wound on Amidon T50-6 toroids and FL4 and FL5 coils are wound on Amidon T50-10 cores.

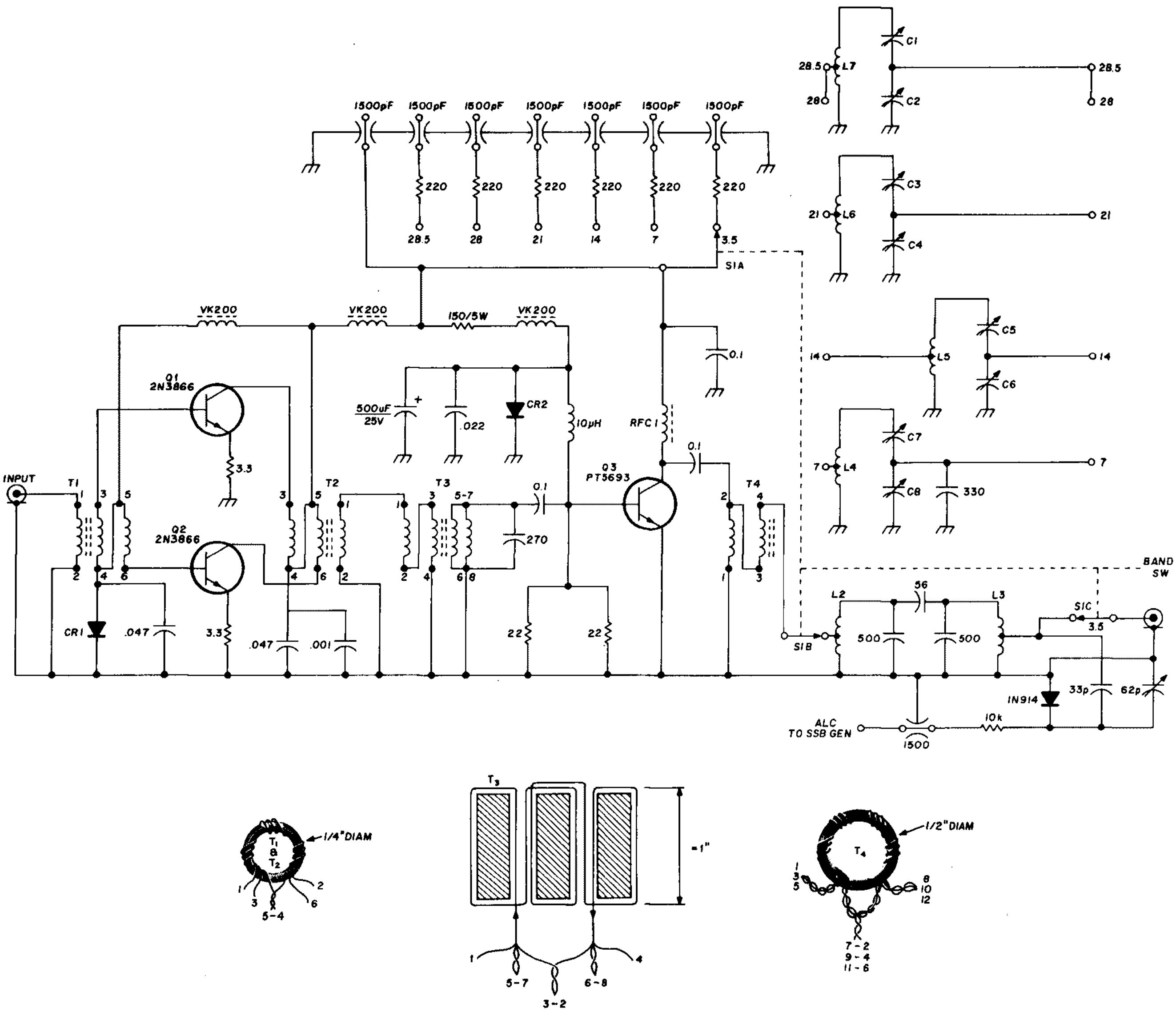



fig. 4. Schematic diagram of the power amplifier. RFC1 is 2 turns of no. 22 AWG (0.6mm) wound on a large Amidon bead. CR1 is a silicon diode rated at 0.4 A and 50 PIV. CR2 is also a silicon diode, 3 A and 50 PIV, stud mounted. T1 and T2 are each 6 turns no. 28 AWG (0.3mm), trifilar wound on a hi-μ 1/4-inch (6.5mm) core (Indiana General CF-2). T3 is a TV balun core wound with four twisted no. 28 AWG (0.3mm) wire. T4 is a high-μ Ferroxcube core wound with 5 turns of 6 twisted no. 28 AWG (0.3mm) wires.

| | capacitors | inductors | | | |
|---|---|---|---|---|---|
| 80 | — | L2 - 3.8 μH, 30 turns no. 24 AWG (0.5mm) tapped 10 turns from the ground end | 15 | C3 - 120 pF C4 - 200 pF | L6 - 0.6 μH, 13 turns no. 20 AWG (0.8mm) tapped 3 turns from the ground end |
| 40 | C7 - 360 pF C8 - 600 pF | L4 - 2 μH, 22 turns no. 24 AWG (0.5mm) tapped 5 turns from the ground end | 10 | C1 - 90 pF C2 - 150 pF | L7 - 0.5 μH, 11 turns no. 20 AWG (0.8mm) tapped 3 turns from the ground end |
| 20 | C5 - 180 pF C6 - 300 pF | L5 - 1 μH, 15 turns no. 20 AWG (0.8mm) tapped 4 turns from the ground end | | | |

Note that L2 through L5 are wound on Amidon T50-6 cores while L6 and L7 are wound on T50-10 cores.

The ARCO trimmers are type 42 or 46 with the values indicated being for resonance. The trimmers, Amidon, and Miller parts are available from Circuit Specialists, Box 3047, Scottsdale, Arizona.

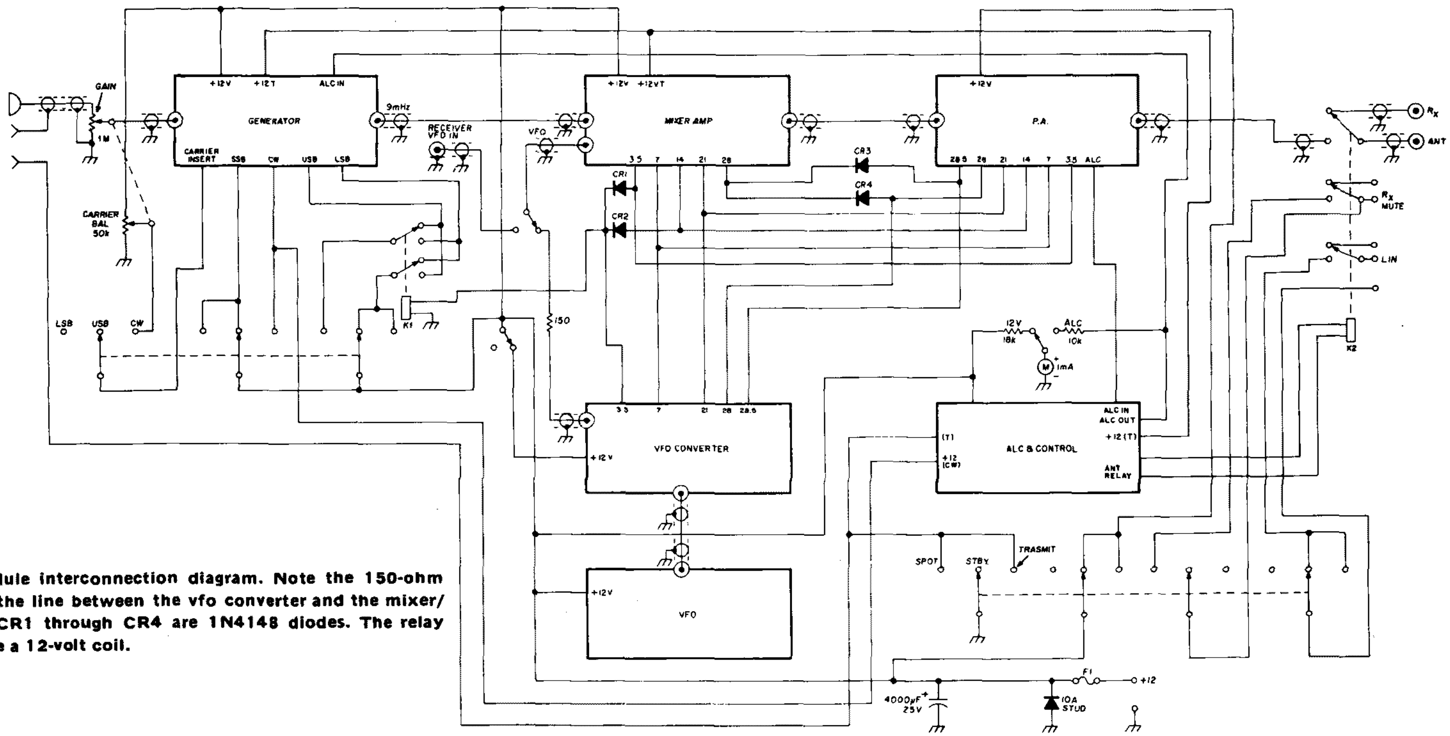



fig. 6. Module interconnection diagram. Note the 150-ohm resistor in the line between the vfo converter and the mixer/amplifier. CR1 through CR4 are 1N4148 diodes. The relay should have a 12-volt coil.

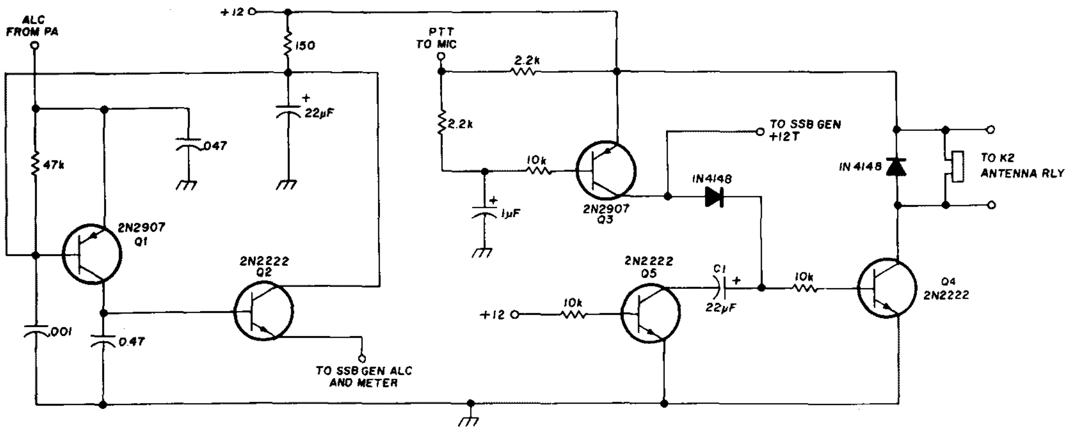
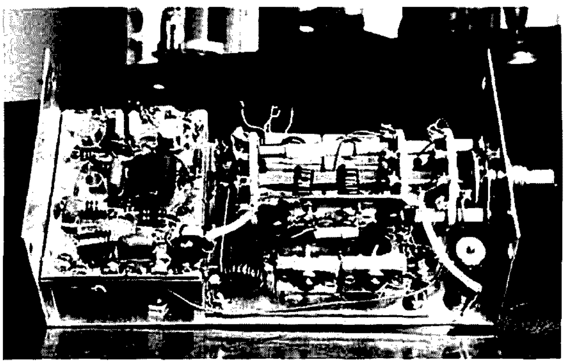



fig. 5. Schematic diagram of the alc and control module.

limits. T1 and T2 are wide-band toroid transformers, while T3 matches the collectors to the base of the final power transistor.

The final power transistor is a TRW PT5693, which was my pride and joy when I first acquired it. It is now a discontinued model but better substitutes are available: PT5649, PT8710, 2N6081 to name a few. The PT5693



Construction of the power amplifier module. The output transistor and associated circuitry is mounted on the heatsink at the left. The large switch selects the correct output network.

is rated at 17½ watts for 175-MHz fm. As a linear amplifier it will provide 10 watts. Forward bias, to improve linearity, is provided by CR2. When rf drive is applied, the necessary increased base current is "robbed" from the diode. CR2 also provides temperature compensation, and should be mounted on the heatsink close to the transistor.

## auxiliary circuits

The alc and receive/transmit control circuits are shown in fig. 5. Q1, with the base and emitter at +12 volts, is normally off. The positive peaks of the rf are filtered and applied to the emitter of Q1. The negative peaks have been clipped by CR3. As the emitter becomes more positive, the transistor starts conducting. The gain of Q3 in the ssb generator is then controlled by this change in conduction level. Transmit control is by Q3, Q4, and Q5. Q3 and Q4 are normally off with K2 de-energized. Pushing the PTT button turns Q3 on, applying +12T power to the ssb generator, turning K2 on. To inhibit K2 opening up between characters on CW, capacitor C1 introduces a delay. In my case 22 $\mu$F worked fine; it may be varied for different dropout times.

## checkout and alignment

I suggest you check each stage individually. As a minimum you'll need a vom or vtvm with an rf probe, a receiver that will tune around 9 MHz, and a grid-dip oscillator. An oscilloscope and frequency counter will help, and of course, a source of 12 volts at 3 amperes.

Starting with the ssb generator, apply power to Q4 and verify there is about 2 volts of rf on the secondary of T2; adjust the core for maximum. Switch to CW, add power to Q3 at +12T and connect an oscilloscope to the secondary of T1. It should indicate about 100 millivolts; again adjust the core for maximum voltage.

Connect the grid-dip oscillator through a 1 to 2 pF "gimmick" to the source of Q5, apply +12 volts and verify that the transistor oscillates with either crystal. With a receiver connected to the 9-MHz output jack, apply power to Q6 and Q7 and check for output. It may be necessary to adjust the turns slightly on L1/L2/L3 for maximum output.

Turn Q4 off, switch to the 9455-kHz crystal and tune a receiver to this frequency. Adjust R2, the 10k pot, for minimum signal. After turning Q4 on again, switch to ssb, and alternate the adjustment of R1 and C1 in the balanced modulator, for minimum signal. In some cases C1 must be connected to the opposite side of the balanced modulator for better suppression. In my transmitter the carrier was down 80 dB.

With the ssb generator connected to the mixer amplifier module, temporarily solder a 47-ohm resistor on the output of the mixer module. The output of the vfo converter, as measured at the 150-ohm interconnecting resistor, should be approximately 0.5 volts. Switch to CW, set the vfo to 3.5 MHz, and apply power to +12 and +12T. You should measure at least 1.6 volts with an rf probe on the output. It may be necessary to adjust the turn spacing of L1 and L2 for maximum output. Repeat this procedure for all five bands, making any slight adjustment necessary to the respective filters.

The power amplifier is normally adjusted for maximum output. A temporarily-connected rf voltmeter is used to measure the output across a 50-ohm dummy load. The respective trimmers, C1 through C8, are adjusted for each band. At this point the entire transmitter can be interconnected as shown in fig. 6.

Set the carrier balance pot to its midpoint and adjust R3 in the mixer amplifier module for 200 millivolts on the secondary of T5. Change to 28-MHz CW and put a voltmeter in the alc line from the PA module. As the carrier level is increased you should get 10 watts output on all bands. Note also that the alc responds. Now, connect the alc line to the ssb generator. Locate the 62-pF trimmer on the PA module and adjust it for proper level of alc action – it should limit the output to 10 watts. Switch to ssb and, using a two-tone test or your mike, verify that this condition also exists. Depending on your voice quality and language (Italian needs lots of alc) you may need to readjust the alc trimmer. The transmitter is now ready for operation.

## reference

1. P. Moroni, I5TDJ, "Solid-State Communications Receiver," *ham radio*, October, 1975, page 32.

**ham radio**


# the remote base:

## an alternative to repeaters

Recommended reading for those wishing to relieve congestion on the vhf bands — a definitive description of the difference between remote-base and repeater stations

We feel that the case for remote base stations, as opposed to repeaters, is a very strong one for those interested in the advancement of vhf/uhf amateur communications. In this article we discuss the remote base-station concept with emphasis on its advantages over repeaters in today's crowded vhf/uhf spectrum.

Appreciable differences exist in the technical details between remote bases and repeaters. The former require a far more flexible command and control system than for repeaters, but they are potentially capable of performing many more functions. Furthermore, the remote base is designed and built with the systems approach in mind and with an eye toward modernization and expansion, whereas repeaters tend to be limited to one or two functions and are generally designed as "common-carrier" machines.

### background

Radio amateurs have explored the characteristics of frequency-modulated communications systems since the 1930s, when Edwin H. Armstrong demonstrated the feasibility of this mode of transmission.[1] Initially failing to win acceptance on the hf bands because of the superiority of ssb in spectrum conservation and weak-signal

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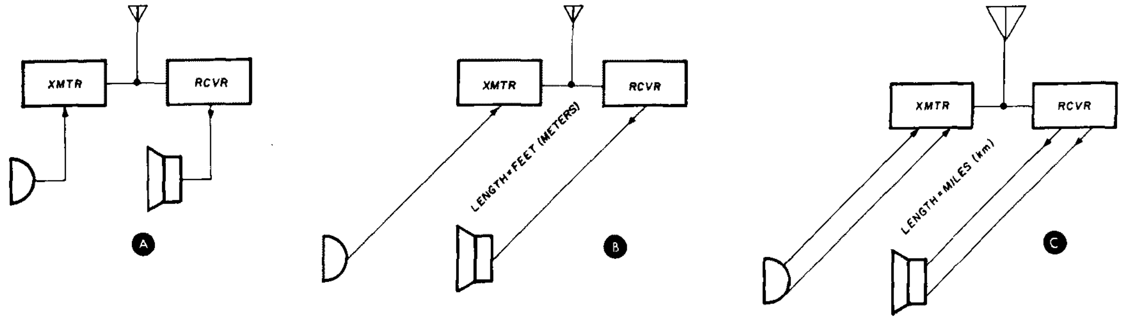
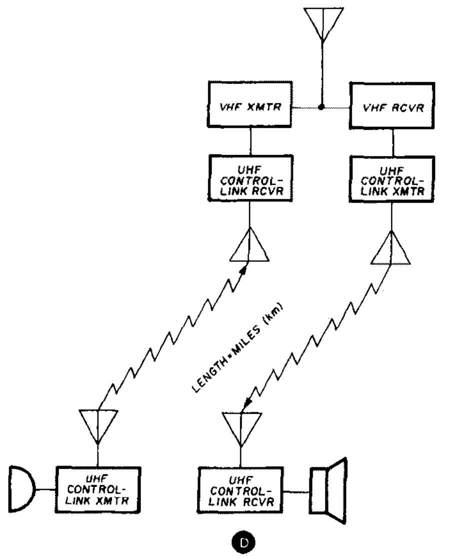



fig. 1. Evolution of a vhf remote-base station. A typical locally controlled amateur station is depicted in A. "Extended" local control is shown in B in which the microphone, speaker and control lines are routed from the operating position to equipment located elsewhere on the premises. A wire-line-controlled remote base is shown in C (transmitter and receiver are located at an elevated site to increase operating range). A radio-controlled remote base station, D, is the same as in C except that the wire link is replaced with a pair of uhf radio channels.




reception, fm entered into general amateur use on the vhf bands in the late 1950s.

Amateurs associated with the commercial* two-way radio business (land mobile service) purchased obsolete police and taxicab radios, retuned them to operate on adjacent amateur vhf bands, and began to experiment with the new mode. Having radios that generally offered one, or at most two, crystal-controlled transmitting and receiving frequencies (or channels), local fm groups quickly adopted standardized channels on which all radios would be operated. In the uncrowded vhf bands of those golden days, these few fm channels were placed well away from existing a-m and CW activity, and the new fm operators were generally ignored. With pretuned radios transmitting and receiving on the same frequency and with effective squelch circuits silencing receiver noise between transmissions, a natural party-line type of operation ensued. Thus the very first amateur fm operations were of a simplex nature — direct, point-to-point transmissions on a single frequency.

In southern California, the first simplex channels were established on 146.760 and 146.940 MHz. Since an a-m repeater (K6MYK) had been in operation at this time, fm operators saw no need for duplication. Instead they concentrated on extending the range of their simplex stations. In the mid 1960s several groups of fm operators established remotely controlled 2-meter fm transmitters on several southern California mountains. These transmitters were operated by radio-control links on the 450-MHz amateur band. Soon thereafter 2-meter fm receivers, tuned to the transmitting frequency, were added to the remotely controlled installations. These early groups of fm experimenters had established base stations (*i.e.*, stations designed to be operated at fixed locations), which were on mountains to increase range. They were remotely controlled and were operated by uhf radio links. These were among the first remotely controlled base stations, or remote bases, as they are more commonly known.

Early remote bases in southern California included those of W6YY, WB6SLR, WB6CZW, WB6LXD, and WB6QEN. From this beginning the number of southern California remote bases has increased to over 100 at present, with a smaller number in northern California, Nevada, and Arizona. Remote bases have been established in other parts of the country though nowhere in the numbers found in California.

## the remote-base concept

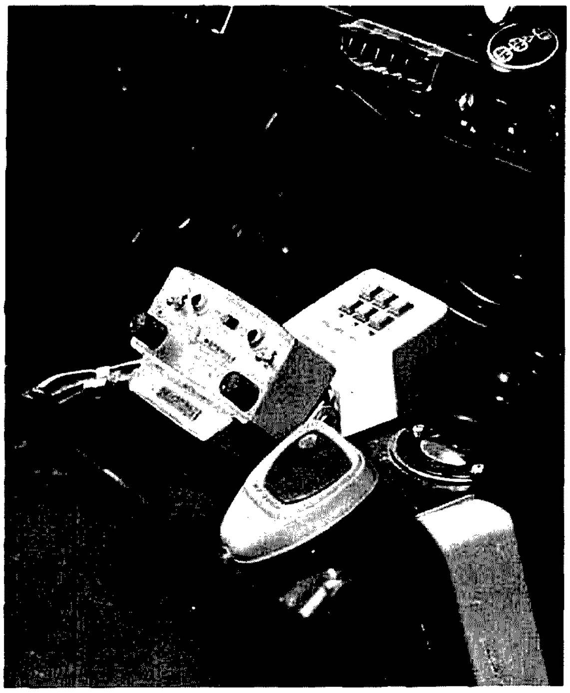
While both remote bases and vhf fm repeaters operate from elevated locations, it should be clearly understood that a remote base is *not* a repeater station. Major differences exist between them in construction, operation, and licensing; these differences are discussed later in greater detail. Most important, however, is the difference in intent of the two stations. Repeaters exist primarily to extend the intracommunity range of user mobile and hand-held portable stations, most operators

*The term "commercial radio" applies to a radio originally designed and manufactured for operation in the commercial two-way Land Mobile Radio Service and adapted to amateur use.

of which are *not* owners or control operators of the repeater. Remote bases, on the other hand, are extensions of the personal stations of their owners and are operated generally only by control operators.

Fig. 1 presents the evolution of the remote base concept. A typical amateur station is depicted in fig. 1A; for the sake of discussion let's assume that it's an fm base station. The owner/control operator talks on the local microphone, listens on the local speaker, and manually turns the transmitter on and off. All controls are at arms' length. This has been the typical style of amateur operation on all bands since the inception of ham radio.

Let's now assume, for reasons of space limitations, that it is inconvenient for the amateur to keep his fm base equipment at his operating position. Since fm stations are operated on crystal-controlled, fixed-tuned



**Several radios may be installed at the remote base and operated through one uhf control station, which avoids duplication of radios between home and car or between several cooperating amateurs.**

channels, it's not necessary to have direct physical access to the transmitter and receiver for tuning purposes. Therefore the amateur may elect to place his base equipment in his basement, attic, or garage, and extend the microphone, speaker, and push-to-talk lines back into his operating position (fig. 1B). Many commercial fm base stations include provisions for doing just this. The amateur is now operating his base station remotely by wire line, although for licensing purposes the station is still under direct control as long as it is entirely contained within the amateur's fixed station license location.

In fig. 1C we extend the operating range of the fm base station by relocating it to a higher elevation. It might be situated at a friend's house on a hill, on a mountain top commercial two-way radio site, or on top of a tall building – all depending on the local geography. The station is still controlled and operated by wire line, but in this case the length of the control line is measured in miles (km) rather than in feet (m). (The technical details of the control system depend on the characteristics of the wire-line pair, its length, and whether or not it is leased from the telephone company). This installation is now a remotely-controlled base station, or remote base, and it must be licensed as a remotely-controlled station. Few, if any, southern California remote bases are wire-line controlled, but the idea has merit for other areas of the country where distances and topography permit.

Now, let's assume that no wire lines can be run to the proposed remote base-station location because of expense, distance, or inaccessibility. It then becomes necessary to control and operate the remote base by radio (fig. 1D). FCC rules, (*Part 97.109a*), require that radio remote-control links operate on frequencies above 220 MHz. While some remote bases operate with 220-MHz radio links and a few others use the amateur microwave bands, the vast majority of remote base operators have elected to control and operate their stations through radio links on the 420-450 MHz amateur band. The reason for this is the availability of high-quality, surplus, commercial fm equipment designed to operate in the 450-470 MHz land mobile service band, or the 406-420 MHz government service band. The former set of radios can be easily retuned to operate in the 440-450-MHz segment of the amateur 3/4-meter band, while the latter set converts easily to the 420-430-MHz segment.

Note from fig. 1D that the control link must be bidirectional. Speech and control information is sent from the local uhf control-link transmitter to the remote base uhf control-link receiver. The information is demodulated and used to operate the vhf fm transmitter. Signals received by the remote base vhf receiver are used to modulate the remote uhf control-link transmitter and are then recovered by the local control-link uhf receiver. The entire control link could be operated on a single uhf channel but this is technically cumbersome. It has become customary to use separate channels for the uhf uplink and downlink. Spacing between the two control-link channels is typically on the order of 5 MHz, a separation sufficient to allow *all* uhf receivers to function properly while their associated transmitters are operating. Thus full two-way duplex operation of the control link is permitted; the control operator can simultaneously transmit signals to, and receive signals from, the remote-base station.

The locally-controlled fm base station in fig. 1A has now grown to become the radio-controlled remote base station in fig. 1D. Fundamentally, however, the only significant change between the two stations has been the replacement of three pairs of wires by one pair of 450-MHz radio links: the pair connecting the micro-

phone to the transmitter, the pair between the receiver and its speaker, and the push-to-talk line pair.

## remote-base advantages

To this point we've discussed the concept of the radio-controlled remote-base station operating on fm simplex channels. While many southern California remote bases have been established to do just this, the description above is actually a restricted view of the capabilities of remote bases. In point of fact, the existence of the basic radio link and control equipment, together with the physical location of the remotely controlled station, represent a resource that can be developed: radios of any type of emission on any amateur band, from 1800 kHz to 10 GHz or higher can be operated remotely. The remote base, for example, allows operation of high-power transmitters, such as on the 50-MHz band, in areas where TVI is a problem. It allows operation on any amateur band where antennas can't be erected at the control operator's location. It affords improved operation on the hf bands where space for efficient antenna systems may be more easily available at the remote-base site.

A remote base offers the opportunity for a group of amateurs to relocate all their radios at one central point while achieving antenna space advantages on hf and height advantages on vhf/uhf. This relocation includes not only home-station radios but mobiles as well. All may be replaced with one uhf radio per location, thereby saving on duplication of radios among several home station and mobile installations.

Those remote-base stations that operate on the fm simplex channels promote spectrum conservation in several ways. With their extended local operating range, they provide interference-free regional-area communications. This can relieve congestion on the hf phone bands by shifting local-area communications to vhf. Because remote bases operate as simplex stations, each occupies only one vhf channel at a time (*i.e.*, 146.940 MHz) rather than two required by a repeater (*i.e.*, 146.340 and 146.940 MHz). Additionally, by the nature of the remote-base design, a control operator always monitors the channel of operation with a mountaintop receiver before transmitting. Thus activity on the operating channel over the entire remote base transmitting range can be easily detected and inadvertent interference avoided. The same is true for repeaters only when a separate receiver and auxiliary link system is used to monitor the output frequency from the repeater site.

Finally, a remote base usually represents the desire of a group of active vhf/uhf amateurs to build a communications system. In deciding to build a remote base, the constructing group does not require the use of the limited set of 2-meter repeater channels. This translates to spectrum conservation. In the southern California area it would be impossible to fit more than the one-hundred remote-base groups into individual 2-meter repeater pairs, even when using 15-kHz channel spacing and all the simplex channels. While it's true that each remote base requires a pair of dedicated channels, these channels are in the spacious 440-450 MHz region. On a narrow-band deviation (±5 kHz) basis, this region of the spectrum contains a potential 200 pairs of channels, with another 200 pairs in reserve between 420 and 430 MHz.

## constructing a remote base

Occasionally an individual will undertake the entire job of designing, building, and installing a remote base. He will then either operate it as his own station, or may invite his friends to use the remote base as co-control operators. More often, in southern California, at least, a group of individuals will be formed to build and operate the remote base, thereby sharing the financial and technical responsibilities. The following comments, although addressed primarily to the group-ownership case, apply as well to single-owner bases.

**administrative and technical responsibility.** A remote base is a communications system that contains separate but intercommunicating radios. The cost and effort to build and operate a remote base is greater than that required to operate a home station, so careful attention should be given to financial and technical responsibilities. One member of the group should be responsible for handling and reporting finances. Provisions should be made for one owner selling his equity in the remote base in the event he must move out of the area. Provisions also should be made for including new members or owners. Lack of adequate preparation in this area has been an historical source of conflict in many remote base groups.

One individual should be responsible for obtaining the site for the remote base, which should be the first task undertaken and completed. When the site involves rental of space at a commercial two-way radio installation, it has been found best to have a single individual from the group maintain relations with the site owner. One individual will have to arrange for licensing the remote base, whether it is in his name or in that of a club station. Additional non-technical duties that may need to be delegated include a) obtaining supplementary permits (for example, from the Forest Service, Bureau of Land Management, local governmental authority) to operate the station, b) maintaining memberships in regional amateur radio associations, and c) providing for fulfillment of public-service commitments.

Technical responsibilities in establishing a remote base should be divided into design, construction, and installation and maintenance areas. A single individual should have overall responsibility for the design of the entire system, although he may wish to delegate specific design projects to others. Particular attention should be given to interfacing between the various subsystems, such as audio and control signal levels between rf hardware and the control system.

Once original equipment designs are complete, construction of individual components can be delegated to group members. Emphasis should be placed on building for reliability, both in selection of components and in construction practices. One or two individuals should assume the responsibility of tuning the rf hardware, integrating the amateur-constructed subsystems

into the final assembly, and performing on-the-ground checkout.

**Maintenance.** When installation of the remote base is completed, the maintenance team assumes responsibility for continued operation. These people should be equipped with the specialized test equipment (wattmeters, signal generators, frequency and deviation meters) for servicing fm communications systems. Inevitably there will be an initial period of system debugging as various design and subsystem deficiencies become apparent. Frequent trips to repair and service the remote will later taper off to occasional visits for scheduled maintenance. At this point the design team will probably begin work on improved subsystems to be retrofitted into the existing remote base, or perhaps better quality rf hardware will be acquired and put into service. Few remote bases are ever truly "completed."

**Rf hardware.** The remote base typically will consist of commercially manufactured rf hardware and amateur-built control systems. Antennas may be either commercially manufactured or home built. In the selection of transmitter and receiver strips, southern California remote-base groups invariably use late-model commercial equipment. All or partially solid-state equipment is preferred for greater reliability, although high-quality all-tube equipment has performed well at some installations for many years. Receivers should have good sensitivity (fet preamps may be added) as well as excellent rf selectivity and cross-modulation rejection; many busy commercial radio sites contain very heavy rf fields. Vhf receivers and transmitters should be capable

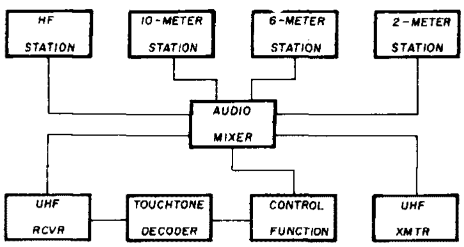



**fig. 2. Block diagram of a typical remote-base station. Control signals from the uhf receiver are decoded in the Touch-Tone decoder, processed in the control-function circuits, and used to operate hf or vhf base station. Speech information is routed to the selected base station through the audio mixer.**

of operation on several different channels, so that the remote base may be switched to operate on whatever channel the control operator wishes to use. The vhf transmitters should be capable of moderate power output (30 - 100 watts), and should be free from spurious output. A remote base operating from an elevated location with a few hundred milliwatts of spurious output will certainly make its presence known.

Commercially manufactured resonant cavities are often used ahead of the entire vhf portion of the remote base to provide additional rf selectivity. The uhf remote base control-link radio should be the best that can be purchased, since it will be the limiting factor in using remote base from distant locations. Matching commercially manufactured 110-Vac power supplies for fm installations are preferred to home-built supplies since they provide the exact voltages required, have provisions for properly interconnecting the transmitter and receiver to other equipment, and are usually rated for continuous-duty operation under severe environmental conditions.

**Antennas.** Antennas and transmission lines for the remote base should be selected with regard to survival under severe weather conditions. Antenna gain, easily obtainable at vhf and uhf is an additional factor to be considered. Remember, however, that many "gain" antennae have major radiation lobes directed at the horizon; for a mountaintop installation it may be preferable to select antennas that radiate their major lobes below the horizon. Transmission lines should exhibit the lowest loss possible; weak received signals and expensively generated vhf and uhf power can be lost in inferior coax. If available, commercially manufactured Foamflex should be used.

**Control systems.** Control systems are the heart of a remote base; they are always amateur constructed. In southern California they vary in complexity from simple audio-tone decoders that drive rotary stepping switches to sophisticated multilevel digital logic circuitry. These advanced systems allow any piece of rf hardware in the remote base to be interconnected with one or more of the remaining pieces in various combinations. Control systems reflect individual needs and capabilities; space prohibits giving specific examples.

A control system performs several functions in addition to enabling transmitters to be turned on and off. In general, the control system must provide for:

1. Authentication and decoding of the received control signals.
2. Selection and activation of the required transmitters and receivers.
3. Selection of specific frequencies to be used within each transmitter and receiver.
4. Processing and conditioning of audio.
5. Automatic indentification of active transmitters.
6. Automatic timing of transmission length to provide ultimate shutdown protection should the control link fail.

Typically, remote bases are controlled by specific audio tones sent along with speech on the uhf uplink channel. The use of *Touch-Tone** audio encoders for this purpose has become relatively standard. The control link usually also contains a subaudible continuous tone squelch signal (*Private Line, Channel Guard,* etc.) as a verification device. Audio-tone decoders, logic circuits, and audio processors are matters of personal preference and design, although some circuits have been published. Timers and IDers are well documented in amateur literature.

**Touch-Tone* is a trademark of American Telephone and Telegraph.

**packaging.** It is considered good construction practice to build all control circuits, timers, audio processors, and identifiers on standard-size edge-connector cards for insertion into a card rack. Interconnections to the individual pieces of rf hardware from the control system are made from the contacts at the rear of the card rack. Provisions should be included in any control system for expansion; the use of individual cards for specific circuits facilitates this goal.

New designs for amateur-built components should be breadboarded and thoroughly tested on the bench before being constructed in final form. In testing, provisions should be made for checkout of the new designs under conditions of continuous duty in temperature and humidity extremes. **Fig. 2** shows a typical remote base station.

One other design feature should be included in any remote base: "series audio." This is illustrated in fig. 3. In a series-audio system, the remote-base vhf receiver runs continuously, even when the control operator transmits; his speech is sent from his 450-MHz control transmitter to the 450-MHz remote base receiver and is then transmitted by the remote base vhf transmitter. The vhf receiver remains on, and although not connected to an antenna during this time, still receives a signal from the vhf transmitter operating nearby. This signal is retransmitted back to the control operator over the 450-MHz downlink. The control operator can listen to his voice as it is being transmitted on vhf by the remote base and can verify that the vhf transmitter in the remote base is being properly modulated. The speech from the control operator follows a path from the control-station microphone back to the control-station loudspeaker, with the remote base vhf transmitter and receiver in "series" with the duplex control link.

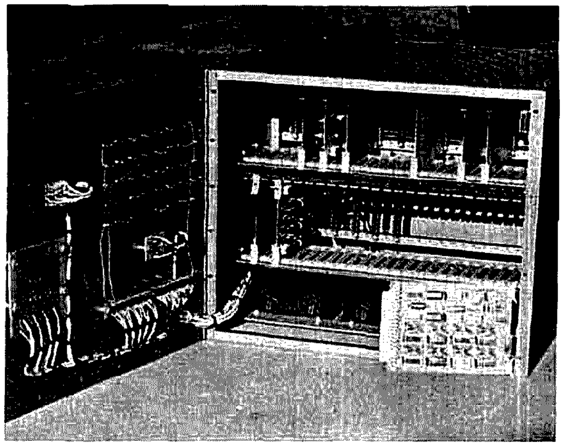
## operating a remote base

What can be done with a remote base is limited only by the imagination and ingenuity of its owners. First and foremost, however, southern California remote bases operate on the area's simplex channels: many can be heard on 146.940 and 146.760 MHz. This is the historical rationale for the establishment of a remote base; and in fulfilling this function, remote bases have helped to remind fm operators — in a time of rapidly expanding numbers of repeaters — that much good work can be accomplished on a point-to-point simplex basis. Occasionally a remote base will be used to transmit bulletins of interest to the regional fm community on 146.940 MHz (a channel that every fm operator can monitor). With their height advantage, many southern California remote bases can be heard from Santa Barbara to the Mexican border; they provide an invaluable resource for tying together an entire region by radio.

Because of the large number of remote bases in southern California, an agreement has been reached that use of the 146.460-MHz simplex channel will be limited to an "intercom" channel among remote bases. This allows two or more remote bases to avoid monopolizing 146.760 or 146.940 MHz, which would prevent mobiles and home-base stations from using these channels. This arrangement has worked well in practice. A number of remote bases also have provisions for operating on 52.525 MHz and 29.600 MHz, the *National* simplex frequencies for these bands.

While it's possible to equip a remote base to transmit on a repeater input channel and listen to the corresponding repeater output channel, such practice is not often done. An exception would be where the repeater to be contacted is so far from a majority of the remote bases' control operators that they couldn't transmit on vhf directly into the repeater from their locations.

Several remote bases have been equipped with hf ssb



**Control system for the remote-base station of WA6ZOI. Card-rack construction is featured. A typical edge-connector card, containing one circuit element, is shown in the lower-right corner.**

transceivers. Notable was the former WA6ZRB remote base, which contained provisions for transmitting on 40 meters including remote tuning of the transceiver vfo. The remote was often used by control operators to check into the WCARS net.

Many remote bases contain autopatches. The use of a remote base for this purpose is particularly fortuitous because it removes the autopatch operation from repeaters in the busy 2-meter band thus reducing congestion and increasing repeater availability for mobile users. Generally, because of nonavailability of telephone lines at the remote base site, a special pair of auxiliary link channels operating in the 420 - 430 MHz region are used to transmit control-link audio from the remote base site to a telephone ground station at a convenient location.

Many remote bases contain auxiliary uhf radios, which link to other remote bases in other areas. Often two or more remote base groups will enter into reciprocal operating agreements, so that by means of the radio links the members of one group, transmitting through their own remote, can control and operate the other remote bases. For example, the Gronk Radio Network can be activated so that stations in southern California can talk to and operate through remote bases in central and nothern California, and in Nevada and Arizona (and vice versa). This is an area where advanced fm operators

are awaiting FCC rules and regulations to catch up to the state of the art.

Several remote bases contain special functions, such as telemetry of prevailing environmental and equipment conditions at the remote base site, or television surveillance of the site.

Remote bases have participated in emergency activities. With their great range and ability to contact virtually any fm-equipped amateur through vhf simplex channels, they provide a natural focus for emergency and disaster operations. Of particular note is the participation of remote bases in the rescue effort after the San Fernando Valley earthquake of 1971.[2] The use of remote bases to relay traffic accidents and other emergencies to public service agencies is a common occurence.

## licensing

Before adoption of Repeater Docket RM 18803, remote bases were routinely licensed by the FCC after the required showings had been submitted. The FCC, then as now, wanted to be convinced that the remotely controlled station would not be tampered with or operated by unauthorized people, and that provisions had been made for automatic shutdown of the transmitters should a failure of the control link occur. Remote-base licenses could be single "additional station" licenses, or primary-station licenses with authorization for remote control. Control operators required no special licenses but were listed as control points on the remote base license.

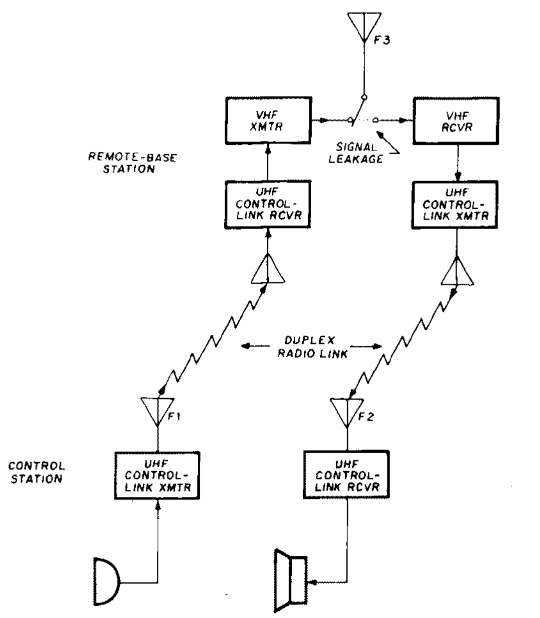



**fig. 3. Remote-base station except for series-audio feature. All transmitters and receivers operate simultaneously. Speech information travels from the control-operator's microphone through the remote base station, then back to the control-station speaker.**

Southern California remote-base operators became concerned with the status of licensing after the adoption of RM 18803. Apparently under the misapprehension that only a handful of remote base licenses would be requested, the FCC devoted its time to the increasing number of repeater applications. But along the way, they released a set of "interpretations" of the new Part 97 rules, which completely changed the nature of remote-base operation.

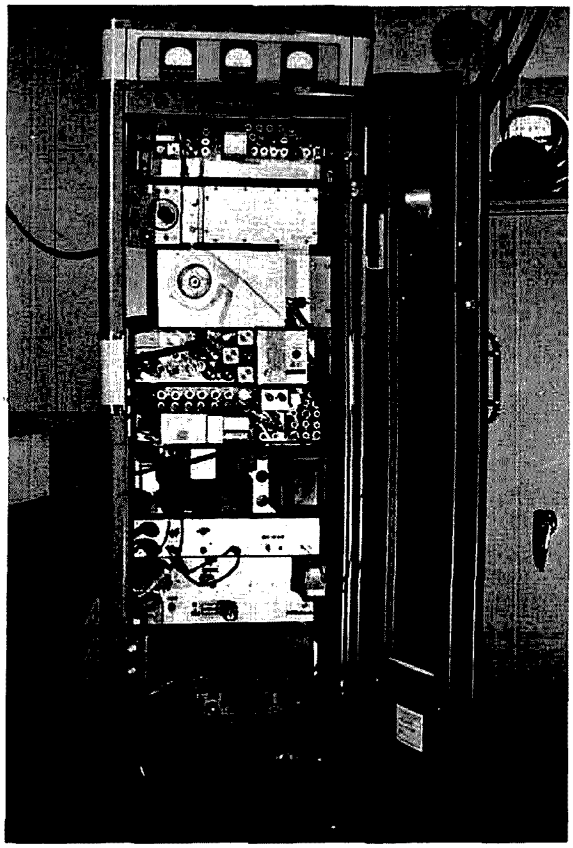
The interpretations included a requirement for a) the licensing as auxiliary-link stations of all uhf transmitters that carry speech to and from the remote base, b) the licensing as control stations of all uhf transmitters sending control information to the remote base, and c) the use of separate uhf frequencies for remote base speech and control uplink channels. A subsidiary effect of these interpretations was to declare as "illegal" the operation of the remote base from portable and mobile locations since, by definition, auxiliary links must operate between two fixed points. The FCC has since dropped the requirement that separate channels must be used for speech and control uplinks.

Nevertheless, remote-base operators are faced with a cumbersome and expensive licensing procedure and with operating restrictions more severe than those before RM 18803. The current licensing procedure, under which the FCC is processing and issuing remote base licenses, is as follows:

The mountaintop remote base must be licensed as a "secondary station" or "club station." This basic license covers the hf and vhf portion of the station; an "auxiliary link" license is required to cover the uhf down-link transmitter. Both privileges may be combined on a single station license for a single application fee. Each control operator must modify his primary station license to include both "control station" and "auxiliary link" privileges; this also can be accomplished with one application fee. The FCC has deleted the requirements for submitting many parts of the required showings, making them instead a required part of the station log.

During the ensuing years the FCC has come to better understand the remote base concept, and has shown increasing willingness to allow remote base (and also repeater) licensees more latitude in the operation of their stations. Docket 21033, which is based in part on a Rule Making petition by the authors of this article, if adopted, will grant essentially complete freedom to operate remote-base stations in the traditional ways described above. For example, Commission restrictions against operation of a remote base from portable and mobile control locations will be eliminated, licensing will be greatly simplified, and the distinction between the remotely controlled base station (with its associated control operators) and the true repeater will be clearly drawn. Southern California remote base operators are generally pleased with the contents of this Docket, and are looking forward to increased flexibility and freedom to innovate.

Remote bases are completely different in intent and operation from repeater stations. Repeaters are operated to extend the communications range for operators of specific mobile and hand-held portable stations interested in communications among themselves. Remote bases are operated as *extensions* of the owners' personal



View of the WA6ZOI remote-base station. Equipment includes a 50-MHz base station, 450-MHz receiver, power supplies and control system. The station is located on Johnstone Peak in southern California.

stations for purposes of communicating with *all* amateur stations. Almost all users of repeaters are *not* control operators, and the act of activating a repeater by transmitting on its input channel is *not* an act of controlling the repeater. Repeater control station operators are responsible for activating the station to repeat the transmissions of other amateurs and for suspending operations in the event that FCC rules are not complied with. By contrast, in southern California, every user of a remote base station has been a control operator. The remote base must be commanded by the control operator through the uhf radio link to operate for each transmission; it is not designed to *automatically* retransmit signals.

The comment has been made that, because the operation of a remote-base station involves speech transferred between hf-vhf and uhf frequencies, the remote base operates as a crossband repeater. From the discussion above it should be clear that the remote-base station does not fit the basic definitions of a repeater. The act of monitoring a vhf channel through the remote-base vhf receiver and uhf downlink channel is not an act of repeater usage. The system *could* be used as a crossband repeater if a) two nonremote base simplex stations were to transmit on a channel being monitored by a remote base, and b) each were to listen to the other through the remote base 450-MHz downlink instead of directly on the vhf channel. In practice this seldom happens; if it should happen it is the responsibility of the remote-base control operator to suspend operations on that vhf channel.

It is our feeling, which is shared by a majority of southern California remote base operators, that liberalization of the present FCC rules (and interpretations of these rules) is required. Ideally each remote base could be licensed as a remotely controlled station, with one license covering the entire mountaintop station including the uhf radio links. Each user would be required to be an authorized control-station operator, having control-station privileges added to his primary station license. There would be no limit to the number of control stations that could be *conveniently* licensed including other remote base stations operating as control stations (many remote bases have 15 control operators at present). The control-station license would confer the privileges of both controlling and operating the remote base and would be usable in portable and mobile operation in addition to its customary fixed-station use. Were these proposed changes to be adopted, remote-base operators would achieve more flexibility to innovate in the amateur vhf and uhf bands.

## conclusion

For those vhf/uhf-oriented groups wishing to expand their interests from individual circuits and individual stations to building an entire communications system, the remote-base concept has several advantages. It offers the chance to experiment with systems engineering – to design and build a system constructed from individual pieces of equipment. The final system can reflect the designers' needs, wishes, and abilities, rather than the standardized requirements of the marketplace. The remote base offers reliable, interference-free local communications capability on the vhf bands, thus helping to relieve congestion on the crowded hf bands. It abolishes the need for duplication of radios between home and car, or duplication among several cooperating owners, and permits the establishment of high-power transmitters and large antennas at the remote-base site. It fosters spectrum conservation on popular bands by removing the requirement for dedicated repeater channels, substituting instead the need for a dedicated pair of channels in the far-less congested 420-450 MHz band. It promotes the use of simplex communications, thus reducing the load of busy repeaters.

Southern California amateurs developed the remote-base concept more than ten years ago. It has proved to be a useful adjunct to the amateur vhf community. We look forward to its adoption in other parts of the country.

## references


1. Thomas McMullen, W1SL, *FM and Repeaters For the Radio Amateur*, ARRL, Newington, Connecticut, 1972, page 7.
2. "The Southern California Earthquake," *QST*, June, 1971, page 60.

# graphical aid for winding rf coils

You'll find this family of parametric curves indispensable when designing small inductors for rf work

Winding a small inductor can be frustrating and usually involves several trials. Here's a simple graphical method that will produce accurate inductance values on the very first try. Carbon composition resistors, ¼ through 2 watts, and standard-size coil forms up to ½ inch (12.5mm) diameter are used for winding the coils.

### literature methods

The usual method for winding coils is to use Wheeler's approximate formula

$$L = \frac{r^2 n^2}{9r + 10l} \tag{1}$$

where $L$ = inductance ($\mu$H)
$r$ = coil radius (inches)
$l$ = coil length (inches)

If all dimensions are in millimeters, eq. 1 becomes

$$L(\mu H) = \frac{0.0394 r^2 n^2}{9r + 10l}$$

These formulas are accurate with one percent for $l > 0.8r$ (*i.e.*, if the coil is not too short).[1] The procedure for winding the coil is described in *The Radio Amateur's Handbook* and several other publications. It usually involves the solution of Wheeler's formula for $n$, searching a wire table for a suitable wire size, then spacing the wire along the coil form to get the required number of turns in the calculated coil length. Also coil inductance slide rules are used such as the ARRL *Type A Lightning*

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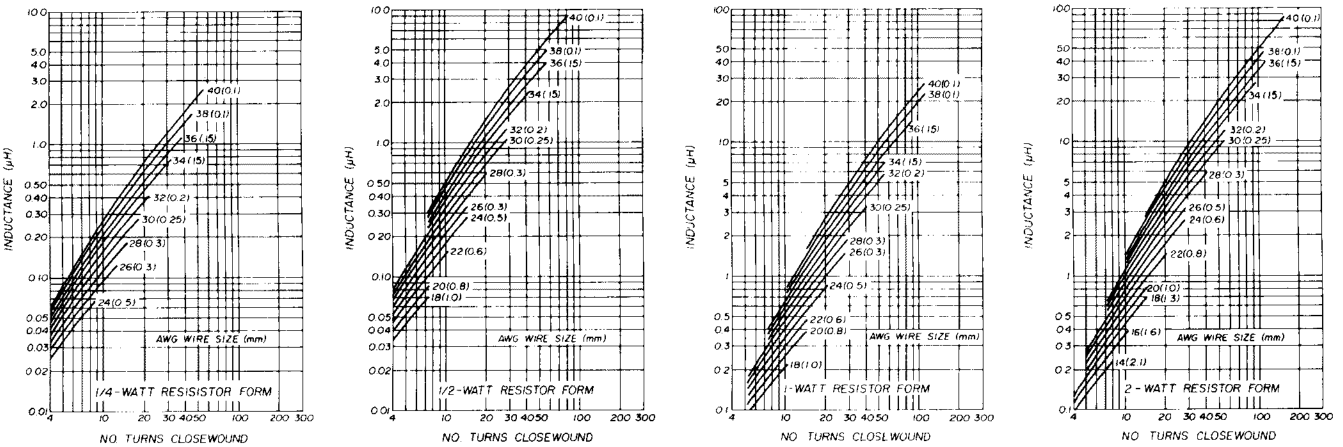


**fig. 1. Curves for finding inductance as a function of turns closewound on composition resistor forms.**

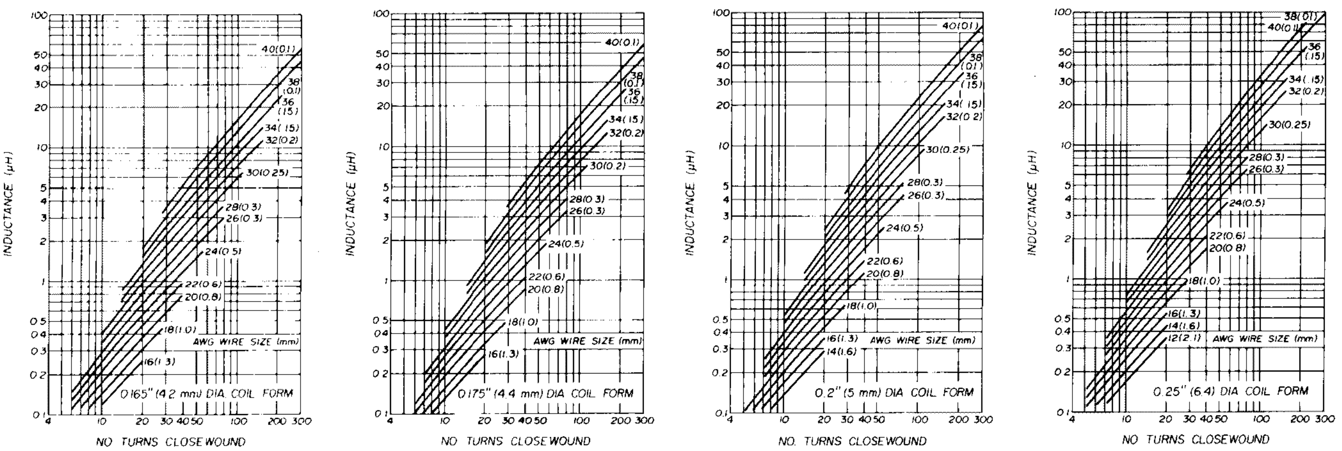


**fig. 2. Curves for finding inductance as a function of turns closewound on standard-size coil forms 0.165 in. through 0.25 in. diameter (4.2 through 6.4mm).**

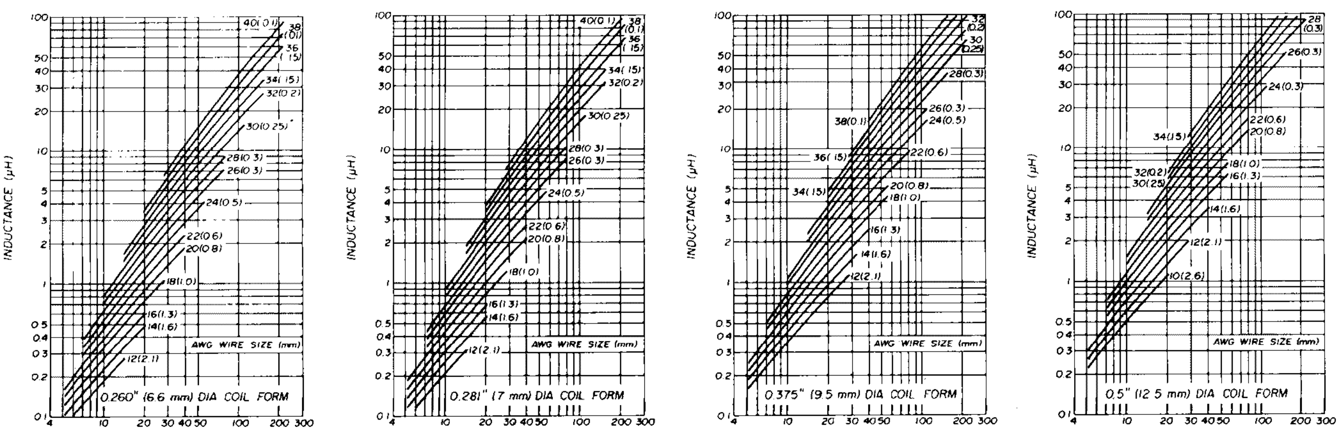


**fig. 3. Curves for finding inductance as a function of turns closewound on standard-size coil forms 0.260 in. through 0.5 in. diameter (6.6 through 12.5mm). A complete set of full-size curves is available from the author for $2.00 postpaid.**

*Calculator*. These work fine for large coil diameters, ½-inch (12.5mm) or greater, but are not calibrated for smaller coil forms.

Accurate small coils can be made if the windings are closewound. Then the wire size determines the coil length (number of turns x wire diameter = coil length). For closewound coils, Wheeler's formula can be written:

$$L = \frac{d^2 n^2}{18d + 40n/T} \tag{2}$$

where $L$ = inductance ($\mu$H)
$d$ = coil diameter (inches)
$n$ = number of turns closewound on coil
$T$= number turns/inch of the particular wire size used for the coil

For dimensions in millimeters, eq. 2 is

$$L\ (\mu H) = \frac{0.0394\ d^2 n^2}{18d + 40n/T}$$

This formula is no simpler than the previous formula, but it is in a form that can be plotted in terms of inductance vs turns for a given wire size and coil diameter. This was done using a Monroe 1666 desktop calculator and plotter. With these graphs and a fair assortment of enameled copper wire, accurate inductors up to approximately 100 $\mu$H can be wound.

## graphical solution

Fig. 1 is used for winding inductors on carbon composition resistors. Composition resistors make excellent forms as their size is standard throughout the industry. The graphs will not let you try to wind more turns on the resistor than it will hold, and the minimum number of turns will be high enough to ensure reasonable accuracy. In winding these inductors keep the following in mind:

1. If the resistor is not going to be removed from the coil use a high resistance value, 100k or more.

2. If the Q of the inductor is important or the frequency of resonance is above 30 MHz, it would be best to remove the resistor form. For very small wire sizes this is impractical as the resultant coil would be too fragile.

3. If the accuracy of the inductor is important, then use as many turns as possible. The greater the number of turns, the more accurate the formula and the graph.

4. When an inductor is to be used as an rf choke, its self-resonant frequency is important, as it exhibits a high impedance at that frequency. Commercially manufactured inductors usually have specified self-resonant frequencies and may be used in a circuit for that reason. So beware — a hand-wound inductor may not work in a particular circuit even though it has the same inductance as a manufactured inductor. To find the self-resonant frequency of an inductor, short its leads and measure the resonant frequency with a grid-dip meter.

Figs. 2 and 3 are used for winding coils on standard-size forms that are normally used for printed circuit work. Larger coil-form sizes of 1/2, 3/8, and 1/4 inch (12.5, 9.5, and 6.5mm) are also included. The same precautions concerning Q and accuracy apply. For these inductors keep the following in mind:

1. If the coil is to be tuned with a ferrite or powdered-iron slug, the frequency used to determine the value of required inductance should be 15 to 20 per cent above the desired value. This is only a rule of thumb and does not take into account the difference in permeability of different slug materials. For brass slugs use a frequency 10 to 15 per cent lower.

2. For slug-tuned coils, the length of the windings should be less than the length of the slug. Typically it should be 75 per cent or less.

3. For air-wound coils, use a frequency 10 per cent below that desired. The coil can then be spread slightly to obtain the desired frequency.

## using the graphs

To gain confidence in these graphs, wind several inductors and check their actual value. A simple way to do this is to make a tuned circuit with a known capacitance then check the resonant frequency with a grid-dip meter. The expression for inductance is:

$$L = \frac{25330}{f^2 C} \tag{3}$$

where $L$ = inductance ($\mu$H)
$f$ = frequency (MHz)
$C$ = capacitance (pF)

I have wound at least one inductor from each graph and find the accuracy better than 10 per cent, which is satisfactory for most home-construction projects.

Example: Suppose a 3.3-$\mu$H inductor is needed for a low-power lowpass circuit and it is desired to keep the size at a minimum. Looking at the graph using a ¼-watt resistor as a form, the maximum inductance is approximately 2.5 $\mu$H, so the next size (or ½ watt) must be used. On the ½-watt resistor graph, the 3.3-$\mu$H inductance line intersects the wire-size lines at 36 turns AWG no. 40, 41 turns AWG no. 38, and 48 turns AWG no. 36 (0.08, 0.10, and 0.13mm). Any of these combinations can be used; however, it would probably be somewhat easier to work with the larger wire size.

For calculating the inductance of toroids see references 2 and 3.

## references


1. Frederick E. Terman, *Radio Engineer's Handbook*, McGraw-Hill Book Company, New York, 1943, page 55.
2. Michael J. Gordon, Jr., WB9FHC, "Calculating the Inductance of Toroids," *ham radio*, February, 1972, page 50.
3. Charles Miller, W3WLX, "Toroidal Coil Inductance," *ham radio*, September, 1975, page 26.


**ham radio**

repair bench

# Bob Stein, W6NBI

## how to use the lab-type rf power meter

When the average amateur hears the words "rf power meter," they bring to mind instruments such as those made by Drake, Swan, Collins, Heath, and others. These are all designed for relatively high powers in the high-frequency and low vhf regions, having full-scale ranges from 20 to 2000 watts. In a more professional class are the through-line instruments manufactured by Bird and Sierra, which use plug-in elements for various frequency ranges between 2 and 1000 MHz, and for full-scale power levels between 1 and 1000 watts.

Power meters such as these are suited for both field and home station use, and require no power source other than the rf energy being measured. However, when it comes to measuring low power, such as the output of a local-oscillator chain, a signal generator, or low-power stages in a solid-state transmitter, they are relatively useless. In such applications, it is necessary to use a laboratory-type power meter which is capable of measurements at the milliwatt and microwatt levels. And, as with most electronic test equipment, a sensitive instrument can be used to make measurements above its range by means of certain auxiliary equipment. Thus, this type of power meter can be a versatile measurement tool for all power levels normally encountered by the experimenting amateur.

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### instrument availability

As with the first article[1] in this series, this one is intended to demonstrate the use of test equipment which is more or less generally available on the surplus market. The instrument which is most often seen is the Hewlett-Packard model 430C Microwave Power Meter. Similar instruments also available from surplus sources are the Sperry Microwave Average Power Meter model 31A1 and the Narda model 440C Microwave Power Meter. Before the word "microwave" turns you non-uhf types off, read on. Microwave is a misnomer, although it must be admitted that these instruments were designed for use at frequencies up to 40 GHz. But they may also be used at 10 MHz and lower; it all depends on the bolometer being used with the power meter.

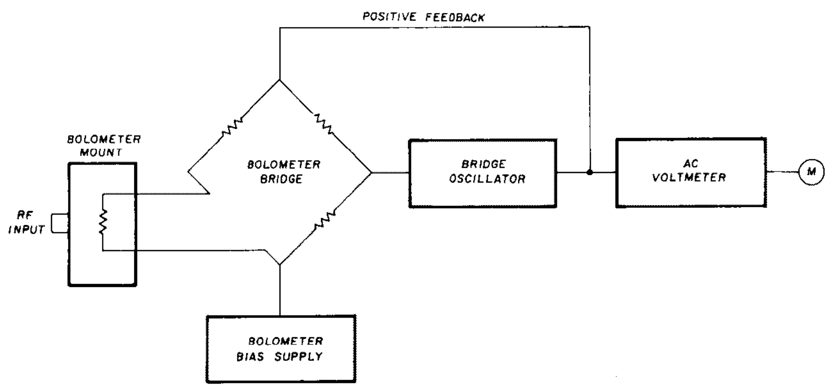



**fig. 1. Simplified block diagram of the basic power meter. The bridge oscillator is tuned to approximately 10 kHz.**

Which brings us to the next point – without a bolometer, a power meter is rather useless. The bolometer is a power-sensing device which is mounted separately from the power meter and connected to it by means of a cable. There are two types of bolometers: *barretters,*

which are normal resistance elements with a positive temperature coefficient, and *thermistors*, which are manufactured from metallic oxide materials which exhibit a negative temperature coefficient. In its simplest form the barretter may be a short length of very fine wire, such as an instrument fuse, or a metallized film resistor.[2]

The three power meters mentioned above can be used with either a thermistor or a barretter mount.* Typical of these (for 50-ohm coaxial systems) are the Hewlett-Packard model 476A Bolometer Mount, which employs barretter fuses, and the model 477B Thermistor Mount. Several others are listed in **table 1**. Obtaining the bolometer mount is generally more difficult than finding a power meter, but they are around. On the other hand, a simple barretter mount suitable for use to over 500 MHz has been built by W6VSV and will, hopefully, be the subject of a forthcoming article.

Later model power meters, such as the Hewlett-Packard model 431 series, are also beginning to show up surplus. This type requires a temperature-compensated thermistor mount, typically a Hewlett-Packard model 478A or 8478B. The instruments may be differentiated by the bolometer connector on the front panel. The older type power meter and bolometer are interconnected via a simple coaxial cable; the newer temperature-compensated thermistor mount connects to the power meter by means of a cable with multi-pin connectors.

The Hewlett-Packard model 430C Microwave Power Meter is commonly available on the surplus market at reasonable prices. It may be used with either a thermistor or a barretter mount.

## how it works

The bolometer-power meter combination functions because of the fact that the bolometer element is essentially not frequency sensitive within its specified range (although there are some variations which are plotted at selected points on the newer mounts). This means that equivalent amounts of power, from dc to the maximum specified frequency, will produce the same resistance in the element.

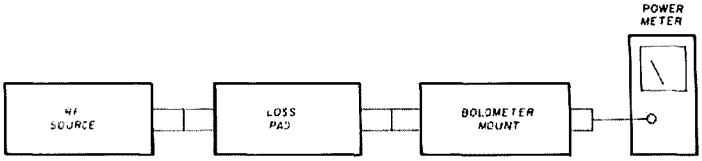



**fig. 2. Using a loss pad to extend the range of the power meter. The pad must be capable of dissipating the power output of the rf source.**

In the earlier power meters, the bolometer forms one leg of a bridge which is connected in the feedback loop of an audio oscillator; see **fig. 1**. Dc bias is also applied to the bolometer element. This configuration results in a self-balancing circuit. The combination of dc bias and audio-frequency power applied to the bolometer causes it to assume a resistance value which balances the bridge. In practice, this balance is achieved by adjusting the bolometer bias controls on the power meter so that the meter indicates zero with no external rf power applied to the bolometer.

When the bolometer is connected to a source of rf power, it heats and its resistance changes, unbalancing the bridge. Since the bridge-oscillator circuit is self-balancing, and the dc bias is fixed when the meter is zeroed, the bridge rebalances itself by reducing the audio oscillator power by an amount equal to the external rf power. This *decrease* in oscillator power is measured by a voltmeter circuit which indicates an equivalent power *increase* representing the external rf power applied to the bolometer.

The Hewlett-Packard model 430C and similar power meters are designed to work with both positive (barretter) and negative (thermistor) temperature-coefficient bolometers which have operating resistances of 100 or 200 ohms. Front-panel switches on the power meter change the bolometer bridge configuration to accommodate these variations.

Because simple bolometers, especially the thermistor type, are extremely sensitive to changes in the ambient

*It appears to be standard in the industry to designate the sensor as either a thermistor mount or a bolometer mount, the latter being a barretter mount. In this article, the two will be differentiated as thermistor or barretter mounts, with the term bolometer used as the overall category which encompasses both types.

temperature, improved thermistor mounts have been developed which are temperature compensated by incorporating additional thermistors in the mount. This requires additional circuitry in the power meter which uses the compensating thermistors to maintain a balanced detection bridge under changing temperature conditions.

## thermistors vs barretters

A comparison of thermistors and barretters applies only to the Hewlett-Packard model 430C, the Sperry Microwave model 31A1, the Narda model 440C, and other similar types which are designed for uncompensated bolometers. (The later, improved power meters all use temperature-compensated thermistor mounts).

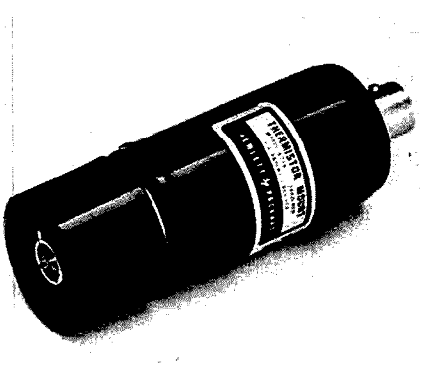
In general, thermistors are more sensitive, have a greater power range, and are less susceptible to overload and burnout than barretters. On the other hand, a barretter responds more quickly because it has a shorter time-constant, and thus is able to follow a modulation envelope better. However, either may be used to measure the *average* power of a modulated signal.

The previously mentioned power meters will operate with either 100- or 200-ohm bolometers. (This refers to the bridge resistance, not the bolometer's input impedance which is usually 50 ohms). Since all of these power meters are able to provide a wide range of bias currents, any bolometer which allows the power meter to be zeroed is suitable for use with that instrument.

Because of the susceptibility of low-level barretters to burn-out, the coarse zero-set control on the power meter must always be turned fully counter clockwise before the bias-current switch is turned on *or* off. This avoids putting a switching transient through the barretter, which might cause it to burn out.

## measuring power below 10 milliwatts

Measuring rf power within the power range of the bolometer and power meter (usually 10 milliwatts maximum) is as simple a procedure as can be imagined. First, connect the bolometer mount to the power meter with an appropriate cable, then set the power meter resistance



**The Hewlett-Packard model 477B Thermistor Mount is a 200-ohm, negatively-temperature-coefficient bolometer. It can be used with the Hewlett-Packard model 430C, the Sperry Microwave 31A1, or the Narda model 440C power meters.**

and polarity switches to the positions which correspond to the bolometer being used. If a temperature-compensated thermistor mount is used, set the calibration-factor switch on the power meter to the factor which is specified on the mount.

Energize the power meter and zero the meter on the power range to be used. If you use an uncompensated thermistor or barretter, this only entails setting the bias-

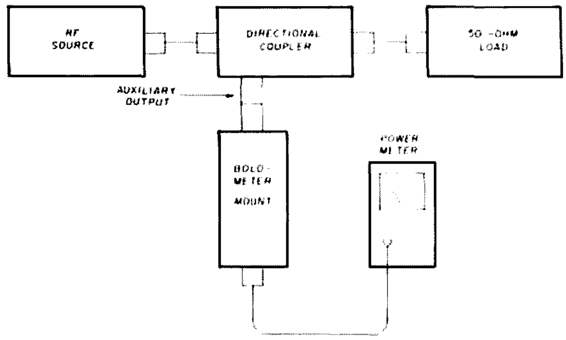



**fig. 3. Using a directional coupler to extend the range of the power meter. The range may be further increased by inserting a loss pad between the auxiliary output port and the bolometer mount.**

current switch to the lowest current range which will allow the meter to be zeroed by means of the zero-set controls. When using a barretter, be sure to observe two precautions:

1. Make sure the coarse zero-set control is turned fully counterclockwise before you turn the bias-current switch, and

2. Do not exceed the maximum safe current specified for the barretter.

When using a temperature-compensated mount, there may be some differences among the various power meters made by different manufacturers. Therefore, the instruction book for the power meter being used should be consulted.

Allow the power meter to warm up. In the case of the older types used with uncompensated bolometers, an hour or more may be required to reach a stable operating temperature, especially on the two lowest power ranges. It is advantageous to have the bolometer mount connected to the device under test during this warm-up period so that the mount and the device under test are at the same temperature.

After the power meter and bolometer have warmed up, re-zero the meter. Turn on the rf power source to be measured and read its output directly from the power meter. Since bolometers have time-constants as high as one or two seconds, you must take this delay into account before reading the meter.

When using the older types of power meters and bolometers, severe drift occurs on the two most sensitive ranges (0.1 and 0.3 milliwatt), even after several hours of warm-up. This drift can be minimized, but by no means eliminated, by protecting the bolometer mount from

drafts and other environmental changes. One way of doing this is to enclose the mount in a block of styrofoam which has been cut and formed to fit closely around the mount. Even wrapping several layers of cloth around the housing will be an improvement over a "bare" mount.

Despite such attempts to stabilize the bolometer, I have found it necessary to re-zero the meter after every reading on the 0.1-milliwatt range. Consequently, I generally take between two and six readings, and average them to arrive at a meaningful measurement.

At this point there may be a question as to why powers of 0.1 milliwatt or less would be of interest to an amateur. A good example of this is checking or calibrating the output level of a signal generator. For instance, there are a great many TS-497/URR signal generators available surplus. Although this military version of the venerable Measurements Corporation model 80 has seen its day, it is quite adequate for the experimenter who is interested in a 2- to 400-MHz instrument. The main problem with such military surplus is that the output level may have to be readjusted. Since the calibration must be made at 50 millivolts output, a sensitive indicating device is required over the frequency range of the signal generator. A power meter and bolometer can do this nicely above 10 MHz, since 50 millivolts across 50 ohms is 50 microwatts, which is half-scale on the 0.1-milliwatt range of the power meter.

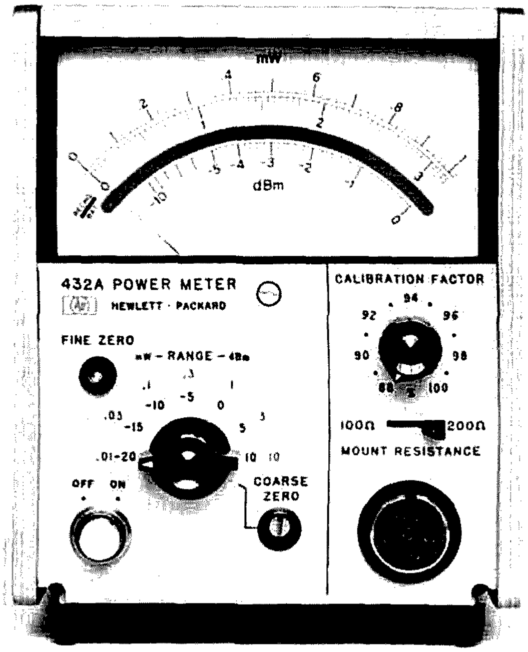
## measuring power above 10 milliwatts

Because most of the power meters and bolometers which are likely to be available have a maximum power limit of 10 milliwatts, signals above that power level must be reduced by some means to use the instruments.

The most convenient method is to introduce attenuation between the rf source and the bolometer in 10-dB steps, which will allow you to multiply the power-meter readings by multiples of 10. Thus, to increase the range to 100 milliwatts, a 10-dB loss pad can be inserted between the source and the bolometer mount, as shown in fig. 2. In the same manner, the power-meter range can be extended to 1 watt by the use of a 20-dB pad, although you must be certain that the attenuator can dissipate 1 watt. The 10-watt level can be reached by using a 30-dB pad, but this must be a power attenuator rated at 10 watts dissipation or greater.

For powers over 1 or 2 watts, a directional coupler will usually be more convenient to use than a loss pad. This arrangement appears in fig. 3. A 50-ohm load, capable of dissipating the output power of the rf source, is connected to the main output port of the directional coupler, which must also be able to handle the full output power. The bolometer mount is connected to the auxiliary output port. Knowing the coupling factor (in dB) of the directional coupler, you need only to multiply the power indication by the power ratio equivalent to the coupling factor to obtain the actual power.

It is also possible to use a combination of both methods to reduce the power to 10 milliwatts or less. For example, let's assume that the power to be measured is expected to be 20 watts, but a 30-dB directional coupler is all that is available. Since 30 dB represents a power ratio of 1000, this would only extend the power-meter range to 10 watts. Thus attenuating the 20-watt input by only 30 dB would result in 20 milliwatts at the auxiliary output port of the coupler, but this can be




The Hewlett-Packard model 432A Power Meter is an improved instrument designed to be used with 100- or 200-ohm temperature-compensated thermistor mounts.

table 1. Frequency ranges and standing-wave ratios of typical bolometer mounts.

| model | type | frequency range and swr |
|---|---|---|
| Hewlett-Packard 476A | barretter | 20 - 500 MHz: less than 1.15<br>10 MHz - 1 GHz: less than 1.25 |
| Hewlett-Packard 477B | thermistor | 50 MHz - 7 GHz: less than 1.3<br>10 MHz - 10 GHz: less than 1.5 |
| Hewlett-Packard 478A | thermistor (temperature compensated) | 10 - 25 MHz: 1.75 maximum<br>25 MHz - 7 GHz: 1.3 maximum<br>7 - 10 GHz: 1.5 maximum |
| FXR N218A | thermistor (temperature compensated) | 10 MHz - 10 GHz: 1.5 maximum |
| Narda 560 | * | 20 MHz - 1.5 GHz: 1.5 maximum |
| Narda 561 | * | 0.5 - 10 GHz: 1.5 maximum |

*May use any one of several bolometer elements, either thermistor or barretter types. Model number is for mount only.

reduced by inserting a low-power loss pad between the coupler and the bolometer mount; a 10-dB pad would increase the overall attenuation to 40 dB, permitting measurements up to 100 watts.

### measurement accuracy

As with all rf measurements, there are many factors which affect the accuracy of the power measurements described. The one of major significance for amateur purposes is the loss caused by a mismatch between the rf source and the bolometer mount. Table 1 shows the frequency range and swr of several mounts. Note that, except for the Hewlett-Packard model 478A between 10 and 25 MHz, all have maximum swrs of 1.5:1 or less. This is a reasonably good load, especially if the source can be tuned to conjugately match the bolometer mount, as evidenced by maximum output power. Of course, using the measurement configurations shown in figs. 2 and 3 reduces the measurement uncertainty to a virtually insignificant figure because of the matching improvement provided by the directional coupler and load and/or the loss pad.

If, however, the bolometer is fed directly from the power source, and the output of the source is fixed, there will be a loss in available power because the load (bolometer mount) impedance will probably not provide a conjugate match to the source impedance. The limits of this loss can be determined from fig. 4, where the solid diagonal lines represent the minimum loss and the broken lines the maximum loss.

As an example, assume that you want to measure the output of an amplifier which has been adjusted for maximum output power into a known 50-ohm load. We can assume then that the output swr of the amplifier (the source) is 1.0:1. Measurements are being made with a bolometer mount (the load) having a maximum specified swr of 1.5:1. The intersection of the 1.5:1 mount-swr line and the 1.0:1 source-swr line lies at

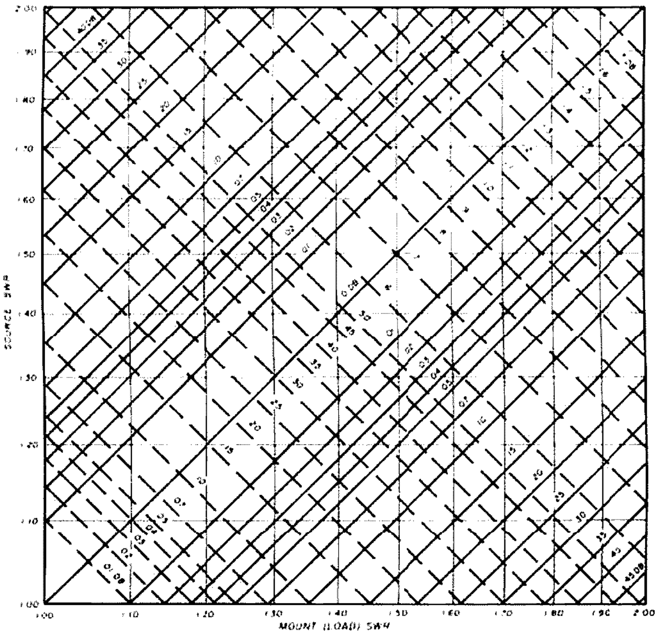



**fig. 4. Mismatch loss as a function of source and load swr. The solid diagonal lines indicate the minimum possible loss, the broken lines the maximum loss.**

approximately 0.17 on the solid diagonal lines and at approximately 0.17 on the broken lines. Since the minimum and maximum losses are equal, there will be a 0.17-dB mismatch loss.

If the same bolometer mount is used to measure the output of a signal generator having a specified output swr of 1.2:1, it can be seen that the intersection of the 1.5:1 mount-swr line and the 1.2:1 source-swr line is at about 0.055 on the solid diagonal lines and 0.37 on the broken lines. Therefore the mismatch loss will be between 0.055 and 0.37 dB.

Other factors which affect measurement accuracy are instrument error, miscellaneous rf loss, dc-to-microwave substitution error, and thermoelectric-effect error. A complete discussion of these errors appears in reference 3, along with a more detailed explanation of mismatch loss.


**The Hewlett-Packard model 478A Thermistor Mount is typical of temperature-compensated units which can be used with the Hewlett-Packard model 431 and 432 series of power meters.**

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3. Operating and Service Manual, Power Meter 431C, Hewlett-Packard Company, Palo Alto, California.
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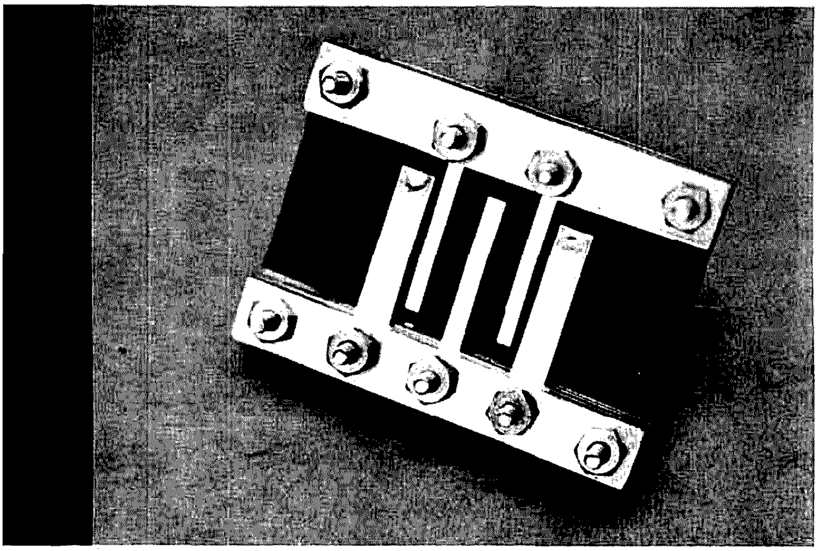


# stripline bandpass filter for 2304 MHz

Interdigital filters can be easily made by using simple hand tools and this stripline design

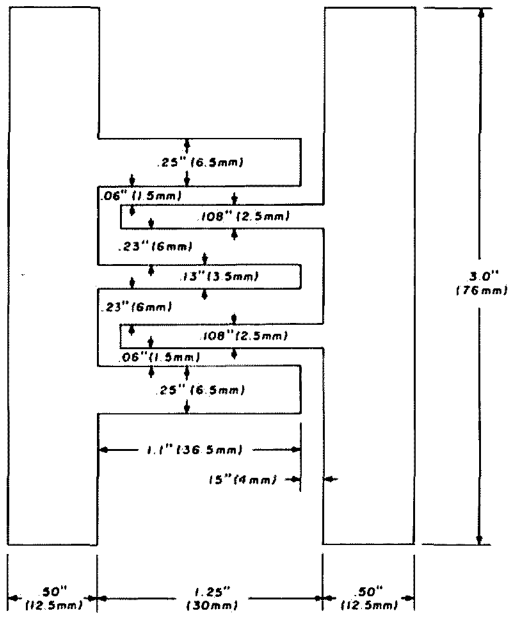



fig. 1. Details of the center conductor measurements and spacings.


By John M. Franke, WA4WDL, Apt. 225, 1006 Westmoreland Avenue, Norfolk, Virginia 23508, and Norman V. Cohen, WB4LJM, 7719 Sheryl Drive, Norfolk, Virginia 23505


Prior to the use of interdigital filters,[1,2] amateurs used only simple half- and quarter-wavelength coaxial-line cavities as filters. Such filters, while easy to construct, lack the needed sharp skirts — interdigital filters have the steep skirts and are only moderately more difficult to construct. However, the need for a lathe to square up the rod ends prompted us to investigate another form of the interdigital filter. An article by John R. Pyle[3]

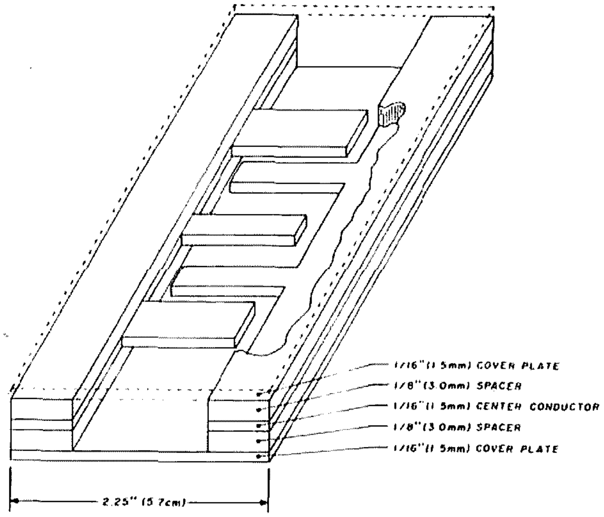
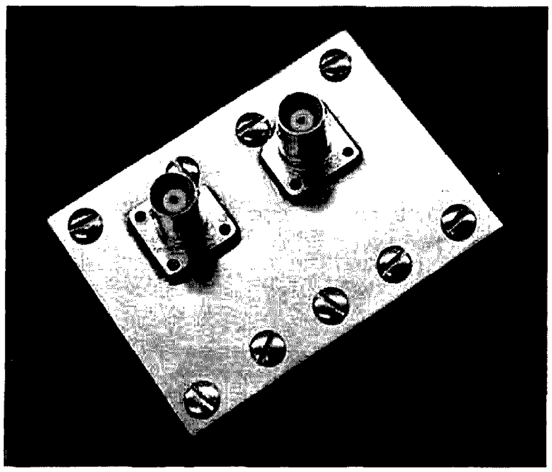



**fig. 2. The sandwich construction of the interdigital filter showing the center conductors and the spacers.**

provided the necessary design curves for fabricating stripline interdigital filters. His paper provided the needed graphs for filters of 1 to 10 percent bandwidth with up to 8 fingers. The curves are normalized for a 1/16-inch (1.5mm) thick center conductor between two ground planes separated by 5/16-inch (8mm).

## filter construction

One such filter recently constructed is shown in the photographs. This filter was designed for 2.5 GHz with a bandwidth of 10 percent. The filter preceded a times-4 multiplier to 10 GHz. The dimensions of the center conductor are shown in **fig. 1**. For other frequencies, the finger widths and spacings are held constant and the fingers are made one quarter-wavelength long. The overall sandwich construction is illustrated in **fig. 2**. The center conductor was made from brass sheet to avoid having to solder the fingers to the root strips. The 1/8-inch (3.2mm) spacers were made from a copper ground strap although brass could be used. The cover plates were fabricated from 1/16-inch (1.6mm) sheet brass. The input and output BNC connectors are sweat soldered to one of the cover plates. The connector center pins were trimmed to just touch the center conductor when assembled, and were soldered to the center conductor before attaching the other cover plate. The sandwich is bolted together with eight 6-32 (M3.5) machine screws. Note that a screw is located near the root of each finger to reduce the contact resistance where there is high circulating current. Alternately, the entire assembly could be sweat soldered together once the fingers have been trimmed. After assembly, the entire unit was given two coats of clear lacquer for corrosion protection. The only tools used in the construction were a hacksaw, an electric drill, and an assortment of hand files.



**Top view of the filter showing the input and output connectors.**

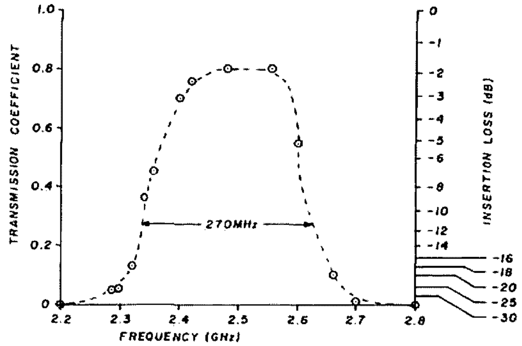



**fig. 3. The measured response of the completed filter.**

The response of the completed filter is shown in **fig. 3**. The midband loss could be reduced by silver plating the unit but we felt this was not worth the added cost. If the center frequency is too high, file away at the spaces between the fingers.

The stripline interdigital filter is a useful alternative to the popular coaxial interdigital filter. The stripline construction is easier for amateurs lacking a lathe and is mechanically stable since the fingers are an integral part of the frame. Future work will involve using double-clad printed circuit board stock for the center conductor to get an even more simple and rugged filter.

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**ham radio**

# the antenna-transmission line analog

## a key to designing and understanding antennas

A practical, non-mathematical discussion of a technique used by professional engineers to design and analyze antenna performance — it is equally applicable to amateur antennas

By **Joseph M. Boyer**, Antenna Consulting Engineer, *17302 Yukon, Suite 63, Torrance, California 90504*

At the present state of the electronics art, the antenna represents the most rewarding, fun-filled, and low-cost area remaining for amateur experimentation. In addition, there is always the challenging possibility of making a real contribution to technology. In all cases, experimentation with radiators will invariably result in improved on-the-air signals.

Many creative amateurs who begin to investigate antennas, however, become frustrated. They have no difficulty understanding certain principles given in elementary treatments of antenna theory, and such initial knowledge carries them through an early fun period of cut and try; but some of the results obtained from such experiments are confusing and demand explanations not found in non-professional books on antennas. If the amateur persists in his experimentation and becomes seriously interested, he finally gets to a point where he wants to know – *before* stringing up more wire or guying up more sections of metal tubing – answers to questions such as, "How do I tailor my antenna design so I can tune over the entire band without the vswr on my feedline climbing to magnitudes into which my rig refuses to load? How much coil reactance does it take to resonate my old 75-meter vertical antenna on the 160-meter band? How efficient is my 20-meter center-loaded whip on the station wagon?" Beyond this, many hams would like to try out their own ideas for antennas but want to know beforehand, with reasonable accuracy, how their brainchild is going to perform on the air.

Giving up on the elementary texts, some of these same amateurs turn to the professional antenna literature for help but are usually stopped cold. Unless they are already engineers by training, they are taken aback by pages covered with the esoteric symbols of the higher mathematics: Fourier series and transforms, Bessel functions, Legendre polynomials, and everywhere copious use of the integral and differential calculus. Few

Mr. Boyer is a prominent antenna consulting engineer who holds twelve patents in antennas, wave filters, and radar targets. Probably best known to amateurs is his low-profile DDRR antenna, which was selected by an international board as one of the 100 most significant inventions of 1963. He has served as an expert consultant to all branches of the U.S. armed forces, and to NATO, NASA, and the Institute for Defense Analysis. Mr. Boyer was previously licensed as W8PVL and now is W6UYH.

amateurs want to go back to college in order to pursue a hobby. Even in the communications industry, many engineers feel there is something mysterious and scary about antennas and leave their design to a handful of specialists.

There is, however, a relatively easy way to avoid the need for using high powered mathematics while getting good, workable answers of engineering accuracy to your antenna questions; answers which will permit you to design complex antennas and predict their performance *before* you build them. The key to all this goes by the rather complicated sounding name of *Antenna/ Transmission Line Analogue*, but the only complicated part about it is its name.

What the Transmission Line Analogue does is to permit an on-paper conversion of your own particular antenna or concept into an equivalent rf transmission line. Once done correctly, this analogue key cranks out answers like your own private computer terminal. You might suspect that because the analogue key dispenses with higher mathematics, it must be some inaccurate, slip-shod method shunned by the real professionals. This is far from true — the analogue key method is used daily by working professional antenna engineers to design commercial and military radiators of all types for use in the frequency spectrum from 10 kHz on up. In its most fundamental form it was used by the brilliant antenna theoretician Dr. Schelkunoff[1] in evolving his powerful mode theory of antennas. As you gain familiarity with the analogue key by usage, you will become positively ingenious in figuring out ways to extend its application into the most involved antenna situations.

To really use the analogue key effectively, however, you must first understand how and why it works and where it comes from. The material which follows may lead you back over some familiar territory, but the route is necessary to establish a certain basic way of thinking about antennas.

## primary mode waves on cylindrical antennas

Fig. 1A shows a perfectly straight, center-fed cylindrical conductor magically suspended without support in free space. You may recognize it as the familiar doublet antenna, having a total length of $2h$ and a half length, $h$. Its uniform conductor diameter, $d$, is twice its radius, $a$. For the moment, forget just how long $2h$ or $h$ is supposed to be in electrical degrees at the operating frequency $f$ (hertz). This doublet antenna has two center input terminals labeled A and B. If you could connect a very accurate rf impedance bridge (complete with its own built-in signal source) directly to the antenna input terminals A - B, without disturbing the invisible fields surrounding it in space, the bridge would read out the doublet's input impedance $Z_{in(A,B)}$ in the form of two separate parts, $R_{total}$ and $jX$. The $R_{total}$ part is the real or resistive part of the entire complex impedance $Z_{in(A,B)}$; the X part with the lower-case $j$ complex operator in front of it (as a label to make sure you separate it from the real part) is reactance. The bridge cannot tell you that the resistive part $R_{total}$ is really made up of two separate resistive parts, or

$$R_{total} = R_r + R_l$$

The $R_r$ is a resistance-like term called the radiation resistance which is a measure of how much wave energy is lost from the antenna by radiation per *rf cycle*.

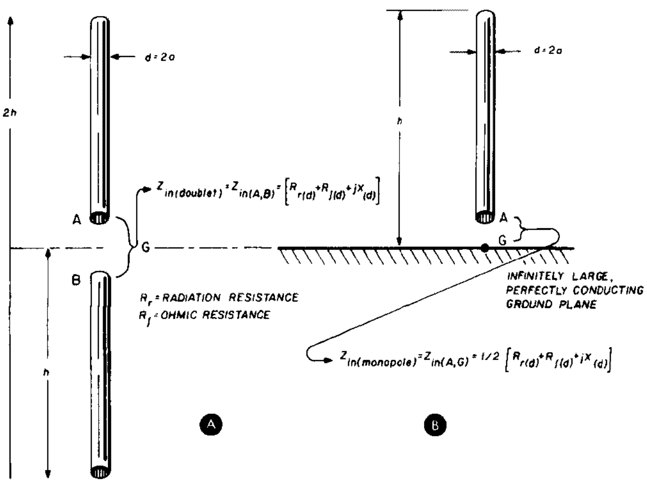


**fig. 1. Center-fed, cylindrical doublet antenna in free space (A), and its feedpoint impedance. Equivalent monopole antenna of equal half-length, $h$, and equal radius, $a$, operated against an infinitely large, perfectly conducting ground plane is shown at (B).**

No one wants the $R_l$ ohmic loss resistance part of $R_{total}$. It just causes some of your input power to terminals A - B to be converted into heat, yet it is always present in real-world antenna elements. One of the battles in antenna design is to keep the ohmic part as small as possible in ways which will be discussed later. In any event, carefully log the input impedance,

$$\begin{aligned} Z_{in(A,B)} &= R_{total(d)} + jX_{(d)} \\ &= R_{r(d)} + R_{l(d)} + jX_{(d)}\ ohms \end{aligned}$$

which you have measured for this particular *doublet* antenna of specific half length $h$, and particular conductor radius $a$, at some exact rf frequency $f$ (hertz).

Now imagine that the half of the antenna connected to input terminal B suddenly disappears, leaving only the other half-length element suspended in free space. A very thin, infinitely large, perfectly conducting metal sheet is then placed exactly through the mid-way point between the former input terminals, the sheet forming a plane lying at a right angle to the remaining antenna half (fig. 1B). The remaining antenna terminal A is now spaced a small distance above the metal plane. By connecting a ground lug to the metal plate at a point directly below terminal A, you now have a monopole antenna (half antenna) operating over a perfect ground plane. Label the newly installed ground terminal with

the letter G. The remaining half of the former doublet is still as before: Same length $h$, same conductor radius $a$. With the rf bridge reconnected to the new input terminals A, G take a reading of the monopole input impedance over perfect ground. You find the new rf input impedance to be

$$
\begin{aligned}
Z_{in(m)A,G} &= \tfrac{1}{2}[Z_{in(A,B)}] \\
&= \tfrac{1}{2}[R_{total(d)} + jX_{(d)}] \\
&= \tfrac{1}{2}[R_{r(d)} + R_{l(d)} + jX_{(d)}] \text{ ohms}
\end{aligned}
$$

$$
\begin{aligned}
\text{or,}\quad Z_{in(m)A,G} &= R_{total(m)} + jX_{(m)} \\
&= R_{r(m)} + R_{l(m)} + jX_{(m)} \text{ ohms}
\end{aligned}
$$

As a result of this first experiment, you write yourself a rather formal note, "The complex input impedance $Z_{in(m)A,G} = R_{r(m)} + R_{l(m)} + jX_{(m)}$ of a cylindrical monopole antenna of conductor length $h$ and conductor radius $a$, erected normal to an infinitely large, perfectly conducting ground plane, is exactly *one-half* the complex input impedance $Z_{in(A,B)} = R_{r(d)} + R_{l(d)} + jX_{(d)}$ of a full doublet antenna of identical half-length $h$ and conductor radius $a$ in free space, when both antennas are measured at the same radio frequency."

With this important experiment out of the way, step back some distance from the monopole antenna erected over the perfect ground plane so that you can inspect its entire length, $h$. Now, really using your imagination, assume that you own a very special pair of eye glasses which permit you to actually "see" electric field lines of force $E$, and magnetic field lines of force $H$. Carefully watching the monopole antenna, again turn on the rf generator so that it supplies energy at frequency $f$ to the monopole input terminals A - G. Fig. 2A is an attempt to show what you would "see."

At the instant you closed the switch (t=1), a small expanding surface like a bubble would appear around the antenna base. Its surface would be covered with dotted E lines pointing radially outward from the surface of the monopole conductor element, with each E line gracefully arching over so that it pointed directly down at right angles to the flat surface of the ground plane. At the same instant you would preceive dashed circles of magnetic field lines of H form concentrically around the monopole antenna element, each growing larger in diameter with the passage of time. Let time suddenly freeze at this point so that you can closely inspect the initial wave surface.

The surface of the "bubble" is a wave front. As this is the first wave to be introduced to the monopole antenna, it's called a *precursor*; a sort of "scout wave" sent out to explore the electrical nature of the yet-unknown antenna to determine — at this one particular frequency — just exactly how the waves to follow will have to finally arrange themselves to be in agreement with certain natural laws.

One of these natural laws dictates that the electric lines of force always point precisely at right angles into the surface of a good conductor such as the antenna element, *and* also point precisely into the flat perfectly conducting surface of the groundplane. Another thing: The "antenna" is the total *combination* of the monopole conductor and the ground-plane surface; the wave front or antenna field is not *in* the antenna conductors, but instead fills the space surrounding the monopole conductor element and the ground-plane surface. The antenna conductor monopole element (or each half of the doublet in free space), together with the ground plane, compose the "nature" of the antenna, and are called the antenna *boundaries*. These boundaries are what the first precursor wave is trying to explore, for they *alone* will determine what finally happens later in time.

Now unfreeze time and let the wave front expand and climb higher up the antenna. Again freeze time at t=2. Now you will notice a very interesting effect: In order to span the increasing distance along the arc between the monopole conductor surface and the ground plane, the E lines get longer and longer as they climb up the monopole. You will also notice that the brightness of the E line arcs closest to the base of the antenna are less intense than those stretching over to ground from higher up on the antenna. Conversely, a fixed radial distance from the antenna, the magnetic field circles around the monopole conductor are intensely bright and glowing around the antenna base, but are less bright as they form around higher parts of the monopole conductor. Clearly, electric field intensity is *increasing* with height up the antenna; magnetic field intensity is *decreasing* with height.

Antenna specialists use the ratio of the magnitude of the electric line of force, $E$, to that of the magnetic field line, $H$, (at any point in space) to define what is called wave impedance, $Z_W$. This is comparable to what the electronics engineer merely calls impedance when he is dealing with the ratio of voltage, $V$, to current, $I$, in circuits physically small in terms of the wavelength of rf energy circulating within them. In contrast, antennas are "big" circuits in terms of the operating wavelength. As a consequence, their fields extend out to great distances around the antenna and the field — rather than voltage or current on conductors — is of first importance.

With this idea of wave impedance in mind, a re-examination of the monopole antenna field discloses that the wave impedance equal to $E/H$ must be increasing in the wavefront as it expands higher and higher up the antenna (because $E$ is increasing and $H$ is decreasing). The wave impedance is small in magnitude near the input terminals A - G, but increases steadily with antenna height, $h$.

This brings up one important way of thinking about *all* antennas: Antennas attempt to perform an impedance-matching function, providing an impedance match between their input terminals and that of the surrounding space by using their conductors as wave impedance "transformers." In this picture, space itself is a common "master" transmission line connecting your antenna with every other antenna in the universe. Such a space transmission line has its own characteristic impedance, $Z_s$, and possesses an infinite number of input

and output terminals. For the moment, however, just assume that this space transmission line surrounds your antenna; it wants to accept the rf energy you are putting into the input terminals of the antenna, but will only accept your wave energy as radiation when certain precise *boundary conditions* are met. The antenna tries to accomplish this feat of getting its waves off into space, but must wait to see what the precursor wave "says" after exploring the antenna.

Let time again unfreeze, and watch the wave front expand through t=3 to the moment of truth at t=4. The wave front has been racing (when you permitted it to) up the antenna at almost, but not quite, the speed of light ($3 \times 10^8$ meters/second). Things appear to be going smoothly so far as the precursor wave is concerned, with those antenna boundaries changing in a nice, gentle fashion. Then *crash*! As if it had smashed head-on into a wall, the precursor wave finds the end of the monopole conductor. To the scout wave this is like an electromagnetic explosion. Let's freeze time again just at the instant this explosion occurs. What happened? To find out, we have to take a cut through the antenna field from a point looking directly down on the monopole conductor. Such a field cut is denoted by the dashed line x-y in view t=4 of fig. 2. The downward looking view is seen in fig. 3.

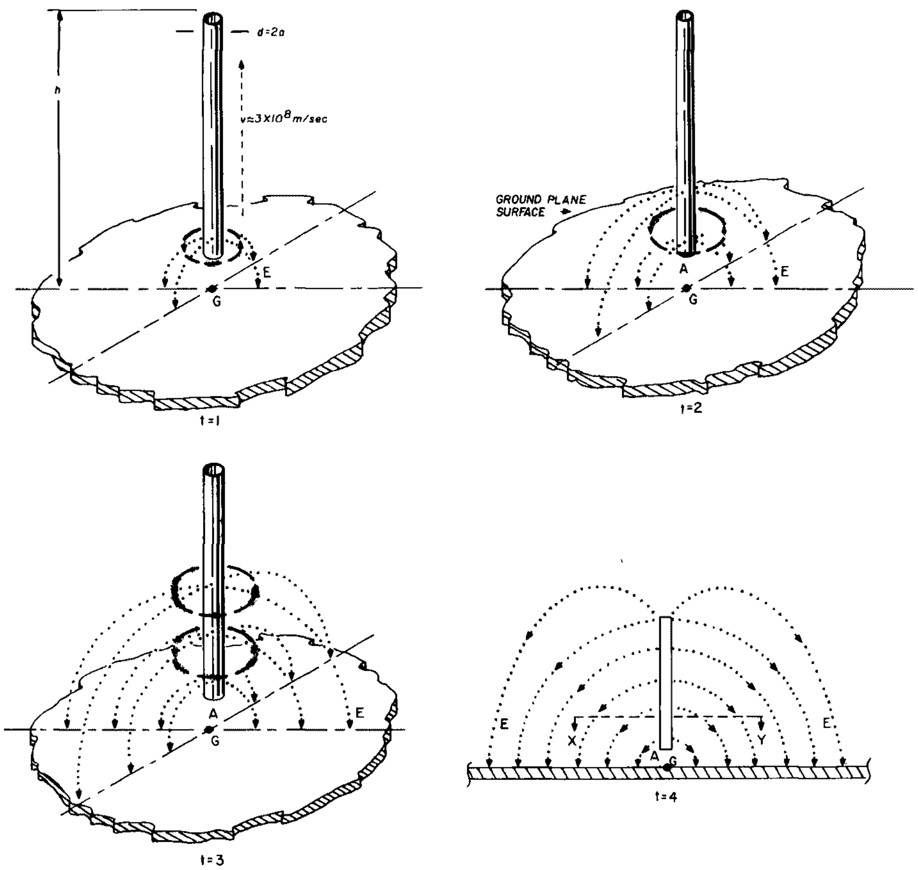


**fig. 2. Precursor TEM wavefront moving up a cylindrical monopole antenna and outward on ground plane G for times t=1, t=2, t=3, and t=4. End reflection occurs at t=4. Top view of the E and H lines at t=4 are shown in fig. 3.**

There are those outward pointing radial electric field lines, $E$, and the closed circles of magnetic lines, $H$. Waves in which the electric field lines lie at right angles to the magnetic field lines entirely in the plane of the wavefront, so that there are no electric or magnetic field components in the direction of wave propagation, are called *transverse electromagnetic* or TEM modes in wave shorthand. **Fig. 3B** shows the wave front inside an ordinary coaxial transmission line with air as an insulator. The similarity of the waves on the antenna and those in the non-radiating coaxial transmission line is no coincidence. Both are type TEM mode waves. In a TEM mode wave, the electric field lines *must end* on the surface of conductors; the monopole antenna conductor and the ground plane serve the same boundary purpose as the inner conductor and inside surface of the shield in the coaxial transmission line. In both cases, the wave fronts are guided in the empty space by the ending of the electric lines onto the oppositely polarized conducting surfaces. The TEM mode waves are held to these boundaries in the same manner as a spider with sticky feet when running around his web.

Now let's try to think the way an antenna theoretician does. Here we have these two classes of waves: TEM mode waves on the antenna in which all electric lines must end on conductors, and space or radiation waves. In free space (say halfway to the planet Mars) there are no electrical conducting surfaces upon which the electric field lines in space waves can end. But we already know that radio waves *can* propagate through free space. That reflected from the open end of the antenna; absolutely no wave energy got away into space as radiation. When a total reflection occurs, the only thing that prevents the space standing wave from reaching a vswr of infinity to one is the very small ohmic resistance of the highly conducting antenna element. Obviously, if that situation continued, antennas, as we know them would not exist. Fortunately, for radio amateurs and antenna men, it

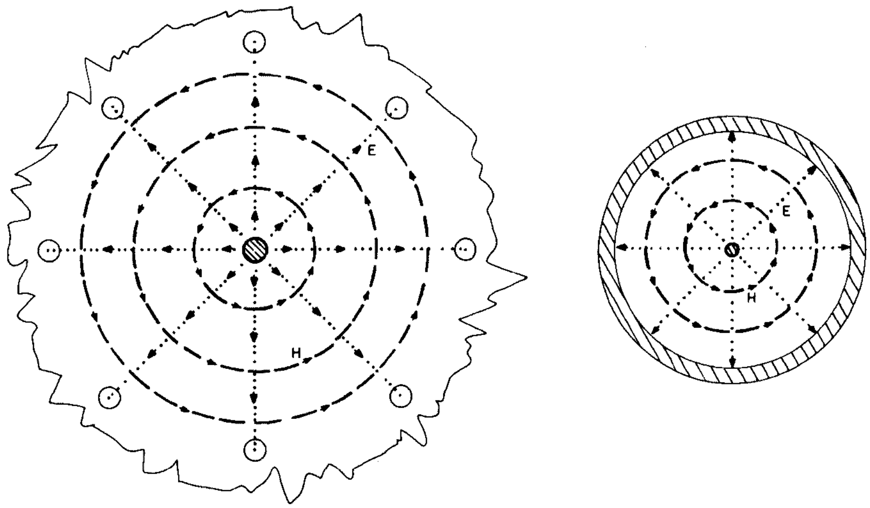


**fig. 3. View looking down on the antenna field at x-y plane indicated in fig. 2. Note similarity to the E and H lines of the wavefront in a coaxial transmission line to right; both fields are type TEM.**

must mean that the kind of waves which can exist as radiation must – regardless of wavefront geometry – contain electric lines which *close on themselves* or form loops the way the magnetic field closes on itself around our antenna. This idea turns out to be correct. There is an infinite variety of space wave modes, but *none* of them includes the TEM mode wave. Therefore, a pure TEM mode *cannot* make the transfer from the antenna boundaries into free space. Free space is an incompatible boundary condition for the TEM mode, precursor wave. Such incompatibility constitutes a huge *impedance mismatch* to the guided TEM mode wave at the *end* of the antenna.

Faced by a large mismatch at the top of the antenna, the TEM mode wave does what all waves do when faced by a mismatch on a transmission line: it is reflected and starts back down the antenna toward the input terminals. In doing so, however, the scout wave encounters other TEM mode waves coming up the antenna in the opposite direction. This kind of situation, with coherent waves moving in opposite directions, always produces the same phenomenon: Standing waves. Note, however, that these standing waves exist in space along the entire length of the antenna and are *not* to be confused with similar standing waves which can form in an antenna feedline because of an impedance mismatch between the antenna input impedance and the line's characteristic impedance.

It should be noted that the TEM scout wave is *totally* doesn't. Nature has arranged things so that as the downward moving scout wave continues to interfere with more and more TEM mode waves coming up the antenna, a wave conversion results; some of the original TEM mode energy is transformed into new, higher order mode waves – wave types which possess closed E and H line geometry and which can make the transfer from the antenna to space. As this converted wave energy (a surprisingly small amount of the total) begins to leave the antenna, energy loss causes the near-to-infinite vswr of the space standing wave to drop to a more reasonable magnitude. The antenna has now reached its steady state of operation.

Here, now, is our antenna: It is "ringing" like a sort of electromagnetic bell as the waves (not charges) race out along the length of the antenna, smash into the top end impedance discontinuity, then race back down the antenna, performing the mode change and supporting the existence of the space standing wave. Each rf cycle produces a small loss of energy to free space as radiation. The actual amount of radiation loss per rf cycle is dependent upon the length of the antenna ($h$ or $2h$) in electrical degrees at the operating frequency, and the conductor geometry (which, for a monopole, includes the ground plane).

Do I hear you say that this picture sounds very much like one describing the way an open-ended (open-circuited) rf transmission line operates? Let's examine that idea! We saw that the wave impedance, $Z_W = E/H$,

was not uniform along the length of the antenna. We also compared a cross section of the antenna field (precursor or scout wave) to the field inside a coaxial transmission line and found them to be the same. Now, even elementary books tell us that, in a lossless transmission line, the ratio of *distributed* series inductance to *distributed* shunt capacitance per unit length solely determines the characteristic impedance of the line as

$$Z_o = \sqrt{L/C}$$

(Advanced textbooks go on to say that the wave impedance, $Z_W$, of the TEM mode wavefront propagating down the transmission line is also a function of this same L/C ratio in the transmission line. Standard types of rf transmission lines, however, have uniform characteristic impedance so therefore they must have a *constant* L to C ratio per unit length.

Such reasoning makes it clear that the cylindrical antenna – when viewed as an rf transmission line – must possess a *variable* ratio of L to C along its length. To reinforce this idea, make another mental experiment: cut out a short section from the monopole antenna conductor. Measure the shunt capacitance to ground of this short conductor section, first at the antenna tip height, then at the antenna midheight, and finally, at the base just above the ground plane. Intuition tells us that the shunt capacitance to ground of the conductor section will be maximum at the antenna base, less at the midheight, and least at the top of the monopole. Fig. 4 shows this same "measurement" result for the case of a doublet antenna. If capacitance to ground (or to the other side of a doublet) varies with position, obviously the L/C ratio cannot be a constant – and that says that the antenna characteristic impedance must also be non-uniform.

But, couldn't we just take this non-uniform characteristic impedance of the antenna and use it as a transmission line model of the antenna in the analogue key method? Yes, but this approach would be a bit messy to put into practical use. Calculations for non-uniform impedance transmission lines are more laborious than those related to lines with uniform impedance. Let's try again. If you have a quantity which changes in some smooth way over a given distance such as $h$, it's possible to take its mean or average value. In high class mathematics, this is called the integral of something (in this case, $Z_o$) over the length, $h$. That is actually what is done: You take this mean or average value of $Z_o$ for the antenna length, $h$, and then use this *average* $Z_o$ as the *uniform* $Z_o$ of your analogue transmission line representing your antenna.

It sounds neat, except that getting this mean $Z_o$ for an antenna is not an easy task. It was solved back in the 1920s at great calculation labor using a dc potential method. Fortunately, later work by Dr. Schelkunoff of Bell Telephone Laboratories has given us some simple formulas to determine the average characteristic impedance of certain kinds of antenna conductor geometry. These conductor geometries include those most often used by amateurs, and professionals alike.

Before presenting these simple formulas, let's make sure we have the concept of the analogue transmission line idea clearly in mind so it may be used with confidence in our experiments on paper with antennas.

## antenna into transmission line

An rf transmission line constructed from extremely high conductivity elements of copper or aluminum, with only air as insulation, represents a very low electrical loss system. Electromagnetic waves moving down such a line stay almost perfectly constant in strength or amplitude even when traveling over long distances in electrical degrees of line length. This constancy of amplitude means that you can use simple trigonometric functions such as the sine, cosine, tangent, or cotangent of the line length in electrical degrees to accurately represent the behavior of waves on low loss line.

On the other hand, if you used poor conductors such as steel or lead to build an rf transmission line, the resistance of the conductors would rob energy from the waves moving down the line and convert it into heat; as a consequence of this energy loss, the wave amplitude would decay or decrease in strength with electrical distance traveled. To represent decaying waves you have to use mathematical functions which also decay in amplitude with electrical distance: Hyperbolic functions. Finally, in correctly representing radiation loss you come up against the cosine and sine integral calculus functions. Not only are these, valuable as they are, a little tacky to use in an amateur technical journal, but I promised at the beginning that only simple math would be needed.

A decision to stick to the use of simple, everyday trig functions means that we must use a lossless equivalent transmission line to represent the antenna. Knowing that real antennas have loss, hopefully the good kind of loss called radiation resistance, how can a lossless transmission line model of the antenna give us accurate answers when solving real antenna problems? Recall the TEM wave mode which did not radiate? That TEM mode would be a uniform amplitude wave representing the major portion of the rf energy oscillating (standing) in the antenna region. If we used only this non-radiating mode wave in the analogue line representing the antenna, the answers would describe *only* the reactive behavior of the antenna at its input terminals. The real or resistive part of $Z_{in}$ would be missing in the answer because it is radiation energy loss which the TEM mode cannot account for in antenna systems. Is that bad? Certainly not! One of the most important things you want to find out when exploring your antenna ideas on paper is how the reactance at the input terminals will change as you move your transmitting frequency over an amateur band, what $jX$ will do if you use a loading coil in the antenna, or how $jX$ changes from one amateur band to another.

The real or resistive part of the antenna's input impedance (which is related to radiation resistance) changes very *slowly* with frequency; the reactive part, however, varies at a much greater rate with changes in operating frequency. *The rate at which the reactive part of antenna input impedance changes with frequency is*

*governed by the antenna's characteristic impedance* when viewed as a transmission line. Does this mean that you just forget all about the real part of $Z_{in}$? No, not at all. You'll end up with a complete answer for $R_A + jX_A$ alright, but you will obtain the real part, $R_A$, the lazy man's way: By looking up the antenna's radiation resistance, $R_r$, as a function of its electrical length $h$ (or $2h$) at the operating frequency, using published graphs of this data. Then you'll add the radiation resistance to the reactive part obtained from the analogue key transmission line key model. It's that simple if a) you are using high conductivity antenna conductors such as copper or aluminum, and b) are feeding the antenna at a current maximum point such as the base of a monopole or the center of a doublet. In the rare case where you are not feeding at a current maximum point on the antenna, then you transfer the $R_r$ value you looked up to the actual feedpoint by a method to be given later.

Incidentally, in deference to Dr. Schelkunoff, it is only fair to mention that in the equivalent transmission line method he evolved, both the real and reactive parts of the total complex input impedance are obtained by using a special lumped "load impedance," determined by separate calculations, which is placed across the end or "output terminals" of the paper analogue transmission line representing the antenna. This more sophisticated

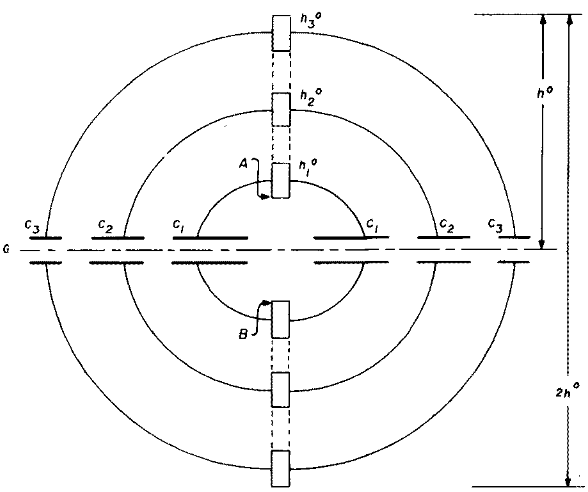


**fig. 4. Variation in shunt capacitance between equal length conductor sections located at the input terminals, mid half-length, and tips of a doublet antenna. Passing the ground plane through G gives the same effect for an equivalent monopole. For a monopole, shunt capacitances double in magnitude.**

technique, however, demands use of advanced forms of mathematics.[2]

## characteristic impedance of cylindrical antennas

Schelkunoff gives the mean characteristic impedance of a doublet antenna with cylindrical conductor elements as,

$$K_A = 120\,(\log_e \frac{2h}{a} - 1)\ ohms \qquad (1)$$

Then, recalling the first experiment where you found that a monopole antenna of length $h$ over a perfect ground plane had an input impedance exactly one-half that of a doublet antenna in free space of half length $h$, and the *same* conductor radius $a$, the mean characteristic impedance of a monopole antenna over ground is

$$K_m = 60\,(\log_e \frac{2h}{a} - 1)\ ohms \qquad (2)$$

Don't let the $\log_e$ part bother you. If your pocket electronic calculator doesn't give the natural $\log_e$ of a number directly, or if you only have tables of the common $\log_{10}$, then

$$\log_e \frac{2(h)}{a} = (2.3026) \times \log_{10} \frac{2(h)}{a}$$

The notation $K_A$ and $K_m$ is used to denote the mean antenna characteristic impedance instead of, say, $Z_{oA}$ or $Z_{om}$. This avoids any confusion with the $Z_o$ of a standard transmission line used to feed the antenna.

## coming up

In the second part of this article I will describe use of the transmission line key method to solve a number of different antenna problems faced by the radio amateur. These will include the design of monopole and doublet antennas capable of being operated over the entire frequency width of an amateur band while keeping the vswr in the feedline down to a specified maximum value into which modern transmitters will load full power. I will also discuss base, center, and higher position coil loading of electrically short monopole and doublet antennas for maximum efficiency. Finally, I will show you how to "dissect" an antenna of your own design into parts to detemine if it will operate as you wish.

Each example will be carefully worked out in full detail (no steps omitted) so you can easily follow the solution and not get lost. In this way you will be able to quickly translate the analogue method to your own problems for any antenna on any amateur band. In the meantime, if you are totally unfamiliar or a bit rusty in the use of elementary plane vectors (phasors) to represent a complex ac impedance, $R + jX$, I suggest you visit the library and get a copy of *Basic Mathematics For Electronics*,[3] or its much earlier version, *Mathematics For Radiomen and Electricians*, by N. Cooke. Cooke was able to teach tens of thousands of Navy gobs to easily master basic ac math on a crash basis during WW II. You will also find him easy to follow and understand.

The difference between an amateur and a professional in a given field of science should not be one of knowledge, but only that the amateur is rewarded in pleasure and the professional in coin of the realm.

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2. S. A. Schelkunoff and H. T. Friis, *Antennas, Theory and Practice*, John Wiley & Sons, New York, 1952, Chapter 13.
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**ham radio**

# novel indicator circuit

## Novel and multi-purpose indicators can be added with a minimum of parts by using the MV5491 red and green LED

Since light-emitting diodes have become available at reasonable prices I find myself using more and more of these devices to monitor the internal circuitry in my equipment from the front panel. LEDs make it easy and economical to do this. I think you gain a better understanding of the internal functions of your equipment through the use of monitor or indicator points. When trouble occurs in the equipment it can often be diagnosed from the front panel with these indicating devices. They can pay their own way both in operator satisfaction and in ease of maintenance, not to mention their esthetic value.

I recently began using a fairly new device, the MV5491 LED manufactured by Monsanto. This device is quite different from the LEDs I had utilized in the past and has proven to be not only interesting, but very practical. Before describing this device I would like to review the standard LED drive circuits I had been using in most of my equipment to acquaint fellow amateurs who, as of yet, have not put these devices to work in their own equipment.

### a standard driver circuit

The standard LED driver circuit that I have been using for the past few years is shown in fig. 1. This configuration uses one-sixth of a TTL hex-inverter buffer, type SN7406, to drive a LED. The SN7406 has an open collector output with the LED and associated current limiting resistor serving as the collector load for the output transistor of the IC. In this circuit, a positive input will provide a low output at the collector, forward biasing the LED and causing it to illuminate. Resistor R1 limits the current through the LED to a safe value as specified by the LED manufacturer. This is a simple, inexpensive circuit and six LEDs can be driven from a

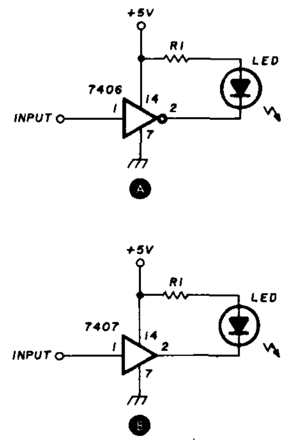


fig. 1. Basic LED drivers configured for normal and inverted inputs.

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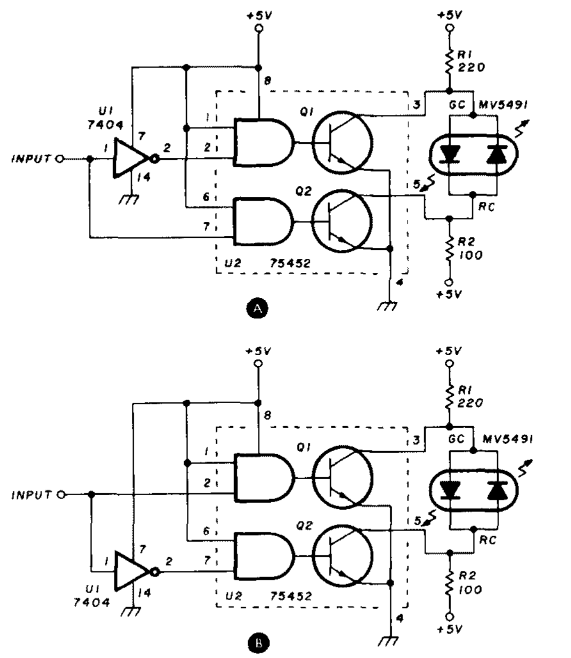


**fig. 2. Two-state LED drivers. The circuit shown in (A) will provide a red indication with a positive input, (B) a green indication.**

single SN7406. This makes for a very low component count and is easily added to existing equipment or new designs. The only calculation required is the value of R1, the current-limiting resistor. To calculate this we must know the voltage and current ratings of the LED. Red LEDs are usually rated at 1.65 volts and about 20 mA. Subtracting the LED voltage and the saturated SN7406 collector voltage from the source voltage, 1.65 + 0.35 from 5.00, we come up with a difference of 3.0 volts which must be dissipated across R1, at the rated current of 20 mA. Using Ohm's law we can calculate the resistor value to be 150 ohms. For the wattage requirement we can use the $I^2R$ formula to arrive at 0.06 watts, so a ¼-watt resistor will suffice.

This circuit works well for source voltages up to about 30 volts, and of course the value of the current limiting resistor must be selected for different source voltages. Most of the LEDs have a maximum current rating of 40 to 50 millamperes and I find that approximately half the rated current will yield adequate light output and assure long life for the indicator. **Fig. 1B** is the same basic circuit but it will light the LED with a low or false input. This is accomplished by using a hex buffer, SN7407, rather than the SN7406. The SN7407 does not contain the inverter function found in the SN7406, but it is pin for pin compatible and the voltage and current ratings are alike. As can be seen in **fig. 1**, you can drive the LED with a true or a false signal, depending on the driver you select. As you get more involved in digital equipment, and I am sure we all will be more involved in this technology in the not-too-distant future, our use of circuits of this type will increase and a working knowledge of them will become more important.

The Monsanto MV5491 is in a standard package like most other LEDs, but is comprised of two light-emitting diodes connected inversely in parallel. One of the parallel diodes is red and the other is green, so steering current in one direction will yield a green light and reversing the current flow will yield a red indication. Stopping the current flow will turn both diodes off, furnishing three distinct indications from one device. The colors are a very vivid red and green, not like some of the earlier single diode devices that were red no matter what the label said!

The red diode section of the MV5491 is rated at 1.65 volts, with a test current of 20 milliamperes and a maximum current of 70 mA. The green diode section is rated at 3.0 volts, with a test current of 20 milliamperes and a maximum current of 50 mA.

The unit comes complete with mounting ring and requires a quarter-inch (6.5mm) diameter mounting hole. When drilling the mounting holes for these and other LEDs in this type package, do not bevel the edges of the holes as this tends to negate the positive effect of the plastic mounting hardware furnished with the LEDs.

## the driver circuit

The project I was working on when I came across this new device used a single five-volt source, so it took a little head scratching to figure out a scheme for driving this new LED. I began scratching out diagrams on the top of the work bench and after an hour or so had an LED that would change color as the input was toggled. I was working on a piece of digital gear so I wanted the driver circuit to be ICs rather than discrete components and eventually came up with the circuit shown in **fig. 2.** This circuit worked out fine and I had an indicator that was green when the input was normal but promptly turned red when an off-normal condition was encountered.

The indicator circuit uses a single SN75452 IC driver and one-sixth of a hex inverter, SN7404, and two

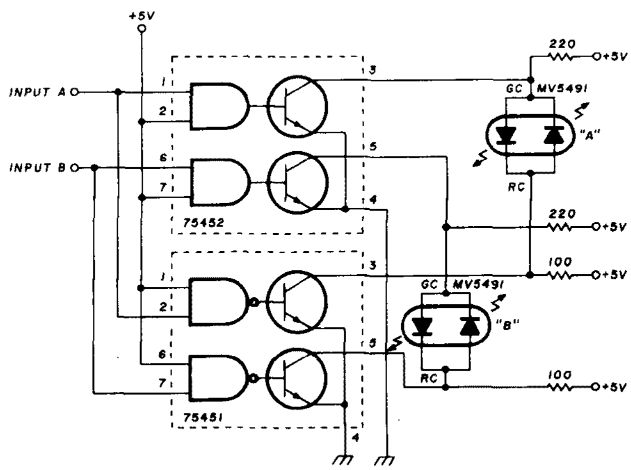


**fig. 3. Two indicators using common drivers and inverters.**

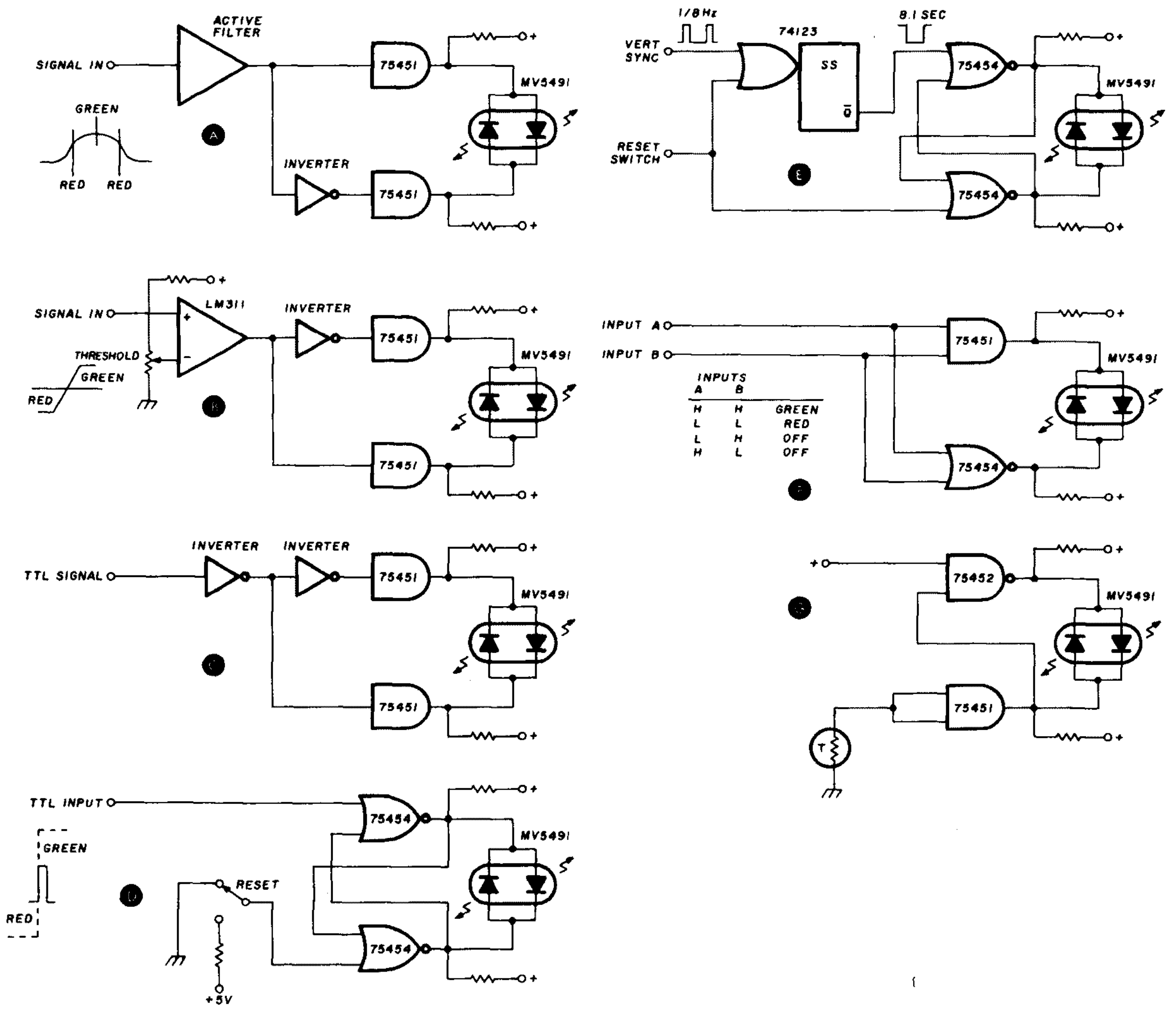


**fig. 4. Applications of the MVS491 LED. (A) shows how the device is used to indicate the presence of a signal in the desired frequency range while (B) uses a comparator to indicate the crossing of a specific level. (C) and (D) can be used as logic probes. (E) is used as a vertical sync indicator for SSTV. (F) is a coincidence detector that indicates equal signals, both high and low, or no signal. (G) is used to indicate when a certain temperature has been reached.**

resistors. It is much like the circuit described in fig. 1, but produces a more profound effect when activated. When this circuit, depicted in fig. 2A, receives an active or positive input, transistor Q2 of U2 is in conduction. This condition brings the red cathode of the LED to ground potential. At the same time, the Q1 section of U2 is cut-off. This brings the green cathode and red anode to a positive potential. With this condition we have the red diode of the LED forward biased (lighted) and the green diode reverse biased (extinguished).

The current through the red diode will be limited by resistor R1. A negative input to the circuit will cause the Q1 section to conduct and the Q2 section to be cutoff. This will forward bias the green diode and reverse bias the red diode. Resistor R2 will limit the current through the green diode. In this configuration, a high or positive input to the circuit will give a red indication and a low input will yield a green indication. If the opposite function is desired, green indication on a positive input, the inverter SN7404, should be placed in series with the Q2 input as per fig. **2B**.

Do not try transposing the LED as the voltage and current specs of the two parallel diodes are different. This is the reason why the current-limiting resistors, R1 and R2, have different values. With the values shown, the red diode is forward biased at approximately 1.65 volts at 15 milliamperes with R1 setting this parameter. The green diode is set at approximately 3 volts and 20 mA by the value of R2. The resistor values are calculated in the same manner as described for the basic circuit of fig. 1, with the exception of resistor wattage ratings.

When Q1 or Q2 is conducting, the entire source voltage is dissipated across the resistors. The maximum dissipation in this case is about ¼ watt, so ½-watt resistors should be installed. In digital circuits where both Q and $\bar{Q}$ signals are available, such as the output of

a flip-flop, the inverter portion of the circuit can be deleted and the out-of-phase signals fed to the driver circuits.

If your application requires more than one indicator circuit, the configuration shown in fig. 3 will reduce the component count. In this configuration two different type drivers are used, one noninverting and one inverting type. These are both in mini-dip packages and will fit in a single dip socket or equivalent space on a pc board. This circuit will drive two MV5491 LEDs and requires only four external resistors. A close inspection of the circuit diagram will show that it uses the internal function of the driver IC to perform the function accomplished by the hex inverter in earlier circuit descriptions. If a number of these dual diode indicators are to be used, this is an excellent choice for a driver circuit.

## applications

I am sure that there are numerous applications for the MV5491 LED, probably as many as there are amateurs still in the homebrew business. I have sketched out a few ideas that I hope to put to use in the future and possibly interest others in their applications.

Fig. 4A uses an active filter preceding the indicator circuit to form a level detector for signals at the desired frequency. This circuit should furnish a green indication on resonance and a red signal at either side of resonance. A circuit or circuits of this type might prove useful for SSTV, RTTY, or subaudio-tone indication for control purposes on fm. This is the same basic circuit described earlier which uses one drive IC and one-sixth of a hex-inverter IC.

Fig. 4B is a take-off on the tuning indicator but would provide an indication with an input change of only a few millivolts, depending on the type of comparator. When used with the LM 311 it will provide excellent performance and can be operated from a single 5-volt power source.

Fig. 4C is a TTL level detector which can be used with TTL, DTL, and RTL logic in the form of a logic probe for trouble-shooting. It will furnish a green indication on a high or plus signal and a red indication on low or false signals. The voltage to operate the circuit can be taken from the equipment under test.

Fig. 4D is a TTL pulse catcher. Again, it can be used as a logic probe for IC circuitry. In this circuit the driver circuit is cross-coupled to form a latch or memory element. A positive-going pulse sets the latch to indicate the presence of a pulse and the latch is reset manually with the reset switch. It should also work well for locating intermittents, such as glitches. You can leave the probe connected to the circuit under test and return to the problem later to see if the latch has been set by an unwanted signal. A circuit such as this one can at times, be more valuable than a scope.

Fig. 4E is a vertical sync indicator for SSTV. In this application we can trigger a retriggerable single-shot with the one-eighth hertz vertical-sync pulse. The absence of a vertical sync pulse will allow the single-shot to time out

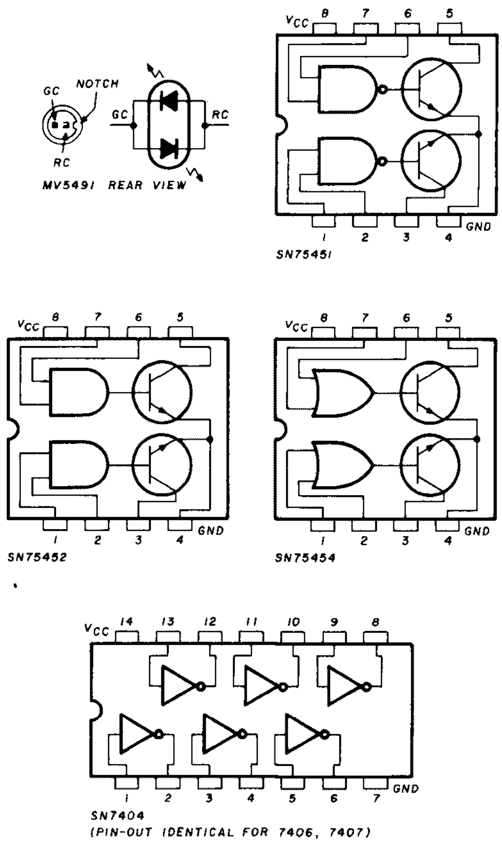


**fig. 5. Device pin-outs for the integrated circuits.**

and change the state of the LED from green to red. The vertical reset switch is then used to restart the vertical sweep and reset the LED to its green state.

Fig. 4F is a coincidence detector with three states. If inputs A and B are both high, the indication will be green. If inputs A and B are both low, the indication will be red. If the inputs are out of phase, one high and one low, the indicator will be in the off or third state. This would provide a good indicator for complex logic circuit monitoring.

Fig. 4G could be used as a temperature-monitoring device with the set points adjusted by a trimming resistor shunted across the thermistor. For photometric application, the thermistor could be replaced by a photocell. A comparator could be utilized before the TTL gate, for increased sensitivity.

I think this is just the beginning for applications of this unique device and I hope this article will stir the interest of other amateurs and further the interest in a return to more homebrew activity. I have enjoyed working with this device and the highlight was the look on a friend's face as the LED changed from red to green. I really had him going until I explained the device to him. This little LED can add the *bells and whistles* touch so often found in today's commercial equipment to your own homebrew efforts.

**ham radio**

# medical data relay via Oscar satellite

## A look at the techniques for the transmission of high-speed data through Oscar

**The Oscar series of satellites** have suddenly provided the amateur radio operator with a whole new field of expertise to learn and experiment with. Now we have reliable and widespread vhf and uhf coverage with minor atmospheric noise. Part of the justification behind the Oscar program is that we, as amateurs, use the satellites to explore and experiment with previously unavailable technologies. Many experiments have involved propagation and its related effects. RTTY, slow-scan TV, mobile operation, and even microcomputer access have all been tried. But, a largely unexplored area involved the use of the satellite to relay data unrelated to amateur radio. Achieving this result could have strong implications in the chosen field.

The practice of medicine today is absorbing new technologies as fast as they appear. This oftentimes carries a requirement that man and medicine be able to efficiently exchange information. It soon became apparent that the data transmission system *I* was developing at the University of Arizona Hospital could be adapted for use through Oscar.

It was originally proposed by 4X4MH to send an electrocardiogram (EKG) through Oscar 6. In the latter part of 1975 this was attempted by W6CG and W7VEW. Success was claimed but nobody knew the parameters necessary to decode the data. It had been an experiment involving "black boxes." At that point it became obvious that the handling of this kind of data was new to amateurs and new techniques had to be learned. I decided to stay with the EKG because it represents a commonly known and needed piece of physiological data. In addition, its bandwidth is wider than most other physiological signals. The ability to successfully relay an EKG would mean that most other types of medical data could be handled as well. Two techniques will be discussed and their merits examined.

To begin with, it's necessary to understand a little about the EKG itself. This is the controlling signal that tells your heart when and how to beat. To satisfactorily represent an EKG for most purposes requires a bandwidth of 0.5 to 50 Hz. Ideally, the lower response should extend to dc. In addition, this signal must be made compatible with amateur ssb transmitters and tape

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recorders if Oscar is to be used in the link. Remember that recorders and transmitters have a lower audio cutoff frequency that may vary from 50 to 300 Hz. Especially in a transmitter, the passband linearity of the audio circuits may vary a few dB. Obviously, the data must be encoded into a new format so that it can be transmitted through the system. Any form of a-m is clearly out. This then leaves fm, which has several desirable characteristics. The first is that amplitude variations due to nonlinear circuits, and a host of other possibilities, disappear. Amateurs experimenting with sstv discovered this in the late 1950s. Second, the use of fm permits the use of a tape recorder to serve as an interim storage device. This is necessary since the source of the EKG was anything but close to the transmitting site. Separate

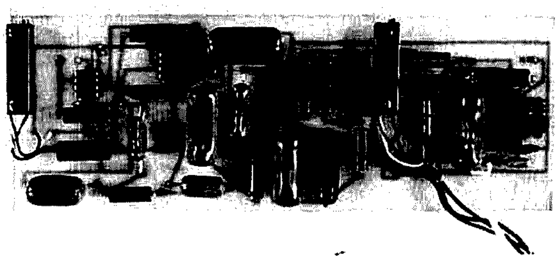


fig. 2. The completed demodulator assembled on a printed-circuit board. The layout of the board is not critical.

scheduling of the various phases of the experiment was then possible. However, the use of a tape recorder introduces some gremlins because the mechanical assembly that moves the tape can introduce errors that are difficult or impossible to eliminate. This and other considerations come into play when designing a system with 1 per cent accuracy.

## analog fm

This approach is the easiest to understand and put to practical use, but it also has its shortcomings. The principle is to use a voltage-controlled oscillator (vco), with its input driven by the data source, as a source of fm. All that remains is to choose an appropriate carrier frequency and set limits on the deviation. At the receiving end, an fm demodulator will transform the varying frequency into a similarly varying voltage. If all goes well, this will provide a faithful reproduction of our data.

## system problems

Since a tape recorder is being used, we must contend with a certain amount of instability in the speed of the tape drive. This speed fluctuation, known as flutter, will modulate the material recorded on the tape and add undesirable components. Rerecording on different machines adds to the problem. If the basic drive speed of the two is slightly different, a dc offset will be introduced. Unless we have a grounded second fm channel available for reference, and the ability to simultaneously transmit the two, the degradation caused by flutter cannot be overcome; it can only be minimized by using quality tape machines and maximizing the deviation of the vco.

It is appropriate now to look ahead and anticipate that Doppler shift and receiver tuning errors will be present. As such, it is unwise to make the vco deviation equal to the maximum that the receiving demodulator can accommodate as there would be no room left for error. With all these factors in mind, I settled on a 1-kHz carrier with ±40 per cent deviation for full-scale input to the vco. The vco I used was a Hewlett-Packard 3310A function generator set to 1 kHz and modulated by the EKG source.

Another problem is noise and here we can only maximize the received signal-to-noise ratio. However, this will not completely delete the effect of noise introduced by the satellite's transponder and atmospherics, especially on 10 meters. Noise causes a directly visible reduction in the resolution of the data.

The receiving demodulator is worthy of a little more attention. In fig. 1, you can see that the audio signal is fed to the input of a 565 phase locked loop. The error voltage of the loop (pin 7) contains the data being

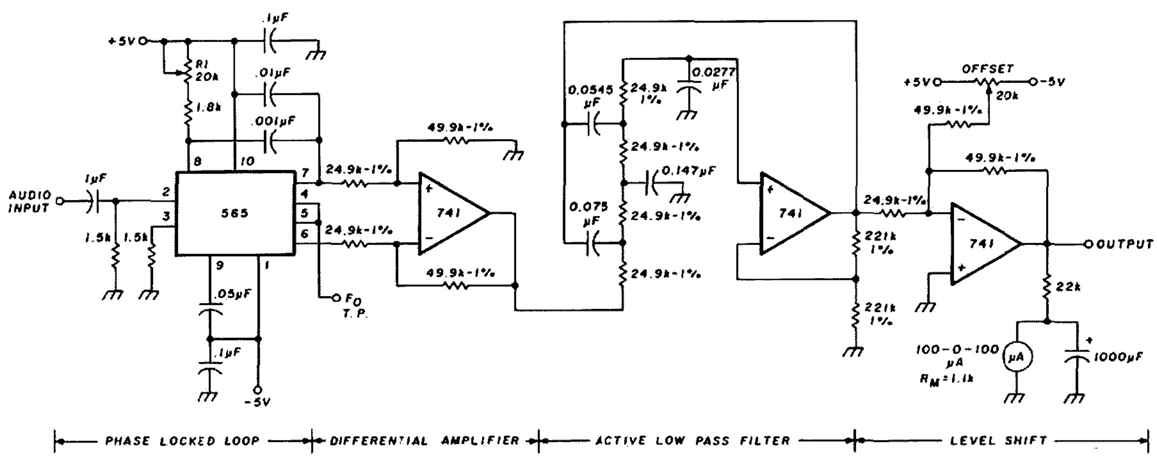



fig. 1. Schematic diagram of the analog fm demodulator. The frequency of the phase-locked loop can be changed by adjusting R1.

sought, along with the undesirable dc and ac components. Following the loop is a differential amplifier. Its purpose is to remove the large intrinsic dc component of the error signal and amplify the rest by a factor of two. Following the amplifier is a four-pole, active-RC, low-pass filter, also with a gain of two. This filter serves to eliminate the high-frequency ac components that are a product of the loop's internal detection process. It also very accurately determines the bandwidth of the demodulator. In this case, the Butterworth filter is scaled for a cutoff frequency of 100 Hz. This response is flat in the passband, down by −3 dB at the cutoff frequency, and rolls off at −6 dB per octave per pole or −24 dB per octave for this filter. The oddball capacitor values are a consequence of an accurate filter design. Labeled capacitor values should not be trusted. I used a pair of 10 per cent capacitors (no ceramics) for each of the four and measured each pair on a bridge to obtain values within one per cent. Lesser accuracy can be tolerated for less accurate results. Following the filter is an amplifier to scale and shift the output to a reasonable value.

## demodulator alignment

To align the demodulator, apply ±5 volts and set the potentiometer to give a loop frequency of 1 kHz at pins 4 and 5. The input should be grounded at this time. Then the potentiometer on the output amplifier is adjusted to give zero output voltage. If the device is now working properly, the loop should lock onto and track any input signal that lies within ±50 to 60 per cent of 1 kHz. The input level can range from 20 mV to 2 volts rms. The output will swing positive for input frequencies above 1 kHz and negative for inputs below 1 kHz. The output-scale factor is approximately 0.5 volt per 10 per cent deviation and is independent of the loop frequency. Note that only one adjustment must be changed to accommodate different carrier frequencies. Fig. 2 shows a photograph of the demodulator.

The 1 per cent resistors used here serve more to guarantee close resistance ratios than absolute values. Readers familiar with op-amp design will note that most of the values in the individual sections can be changed, except for the filter ladder, as long as the resistance ratios are accurately maintained. This ability lends itself to the use of available units rather than pursuing absolute values.

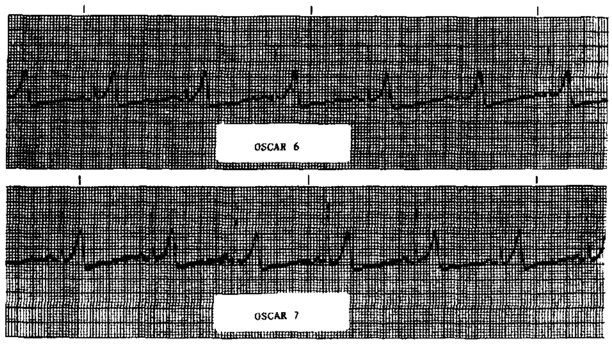



fig. 3. Analog electrocardiograms that have been transmitted through Oscar 6 and 7. Note the less accurate response of the Oscar 7 Mode A version.

This demodulator can also be made to serve as a tuning indicator during the test transmission if you have some prior knowledge of the data. The EKG waveform stays mostly near zero and deviates rapidly when it does so. If a *zero-center microammeter is bypassed with* a capacitor to provide a one to two second time-constant with the meter resistance, the meter will respond to very slow changes such as Doppler shift but will show little response to the faster EKG data. This yields an indicator that tells how to correct the receiver tuning to offset the Doppler shift. All you have to do is keep the meter centered during the transmission by adjusting the receiver tuning. Incorrect tuning will deflect the meter to one side or the other. This method is considerably more accurate than tuning by ear because changes of 10 per cent are quite visible.

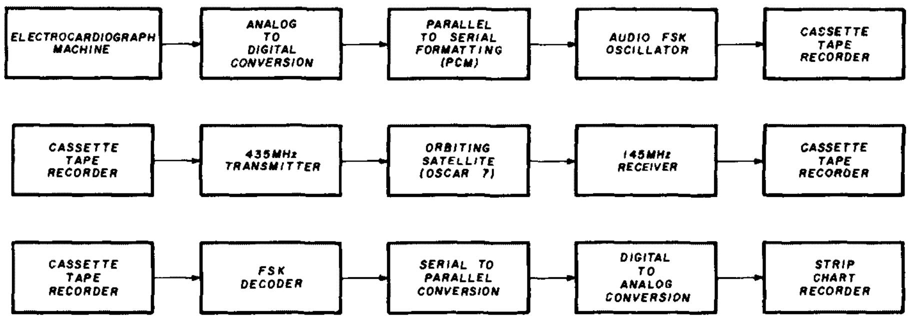



fig. 4. A flow chart for the PCM type system. This method allows the process to be done at different times and locations.

Another method of correcting for Doppler involves the use of a second non-modulated audio carrier sufficiently removed from the first so it doesn't interact. The tuning indicator is locked onto this new carrier. However, this may exceed the available audio bandwidth and requires bandpass filters.

In September, 1975, I conducted a test of this

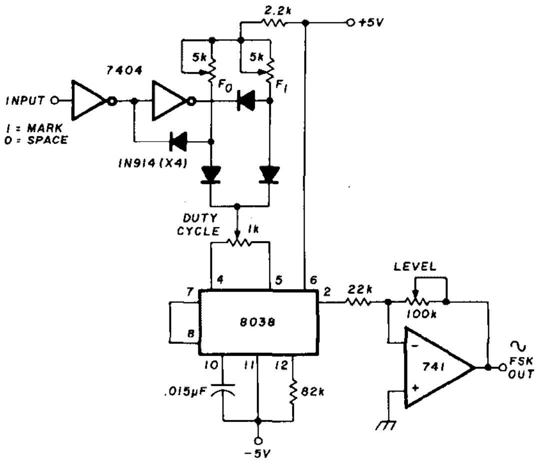



**fig. 5. The fsk oscillator. The separate timing resistors set the mark and space frequencies. A sinusoidal waveform is synthesized internally.**

method with W6ELT transmitting a cassette tape through Oscar 6 and Oscar 7, mode A. Shortly thereafter we tried another transmission to both myself and the radio club at the National Institute of Health in Bethesda, Maryland. The results are shown in fig. 3. Note the difference in the results between Oscar 6 and 7. This is due to the stronger signal available from Oscar 6. Although perfectly copiable signals were present, there is still a residual amount of noise present. This is best seen in the baseline of the EKG. In addition, even though the Doppler was manually compensated for, this method would have been useless if the data had been of a completely unknown form. Obviously, there is room for improvement in many areas.

## going digital

The quickest way to solve many of the problems is to go to a digital method of encoding data. Not only can better resolution be achieved, but the effects of noise, Doppler, and tape flutter can be simultaneously eliminated. However, you must pay the price in terms of system complexity and cost. But if truly accurate, reproducible results are required, this is the way to go. This method is commonly known as pulse code modulation (pcm).

The principles are to convert the data into a digital word and then handle the bits serially, one at a time. Unlike analog techniques, digital signals can be regenerated in the presence of noise. The resolution of the system is determined by the number of bits that represents the data. I chose eight bits which gives a resolution of 0.4 per cent of full scale. This method of handling the data contains a number of parallels to RTTY and some of the techniques can be directly applied. The conversion from analog to digital is accomplished by a commercial module appropriately called an analog-to-digital (A/D) converter. For the purposes of this discussion it is not necessary to understand the inner workings of this and other modules used. It is only necessary to understand their function.

Following the conversion to digital, the parallel eight-bit word presented by the A/D converter is transformed into a serial bit stream by a rather amazing IC called a universal asynchronous receiver transmitter, or UAR/T. This IC consists of two halves. The transmitter

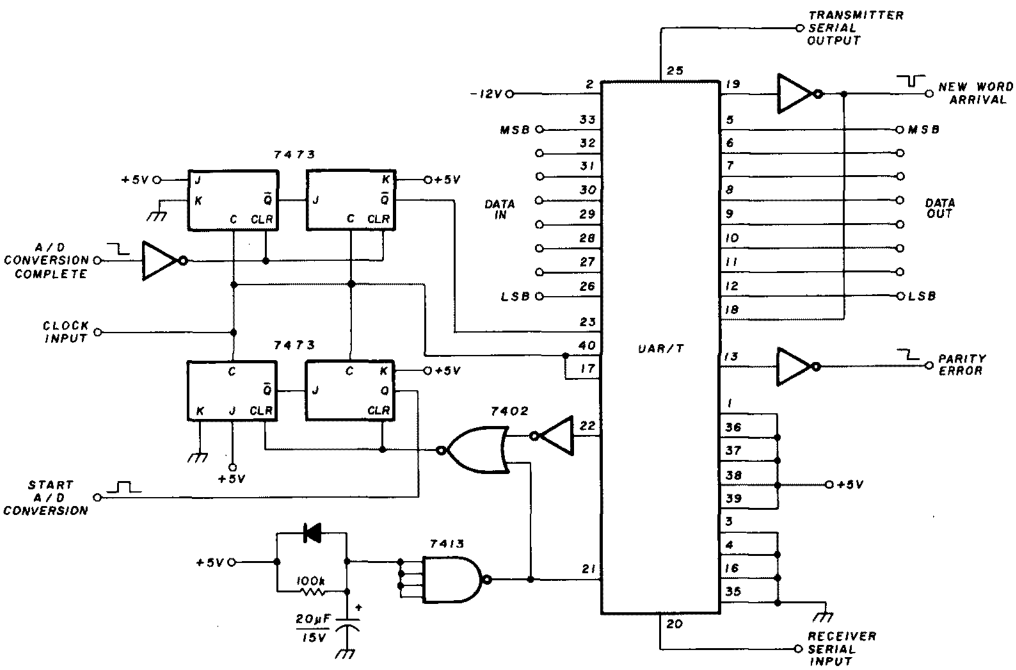



**fig. 6. Schematic diagram of the UAR/T data conversion system. This method performs all conversions between serial and parallel data.**

half takes a parallel input and formats it into a serial asynchronous code with start, stop, and parity bits added to the data. The receiving half does just the opposite. It accepts a serial input, strips off the bookkeeping bits, checks for errors, and outputs the digital word in parallel. Controlling all of these operations is an external clock which determines how fast the system runs. Except for the clock, all of this is nicely packaged into a single 40-pin IC.

The serial-bit stream from the UAR/T then controls an fsk oscillator. This oscillator is switched between two discrete audio frequencies and is an audio representation of the digital signal. These audio signals are then placed clocking out this word serially immediately. As soon as the UAR/T is ready to receive another word, it signals by setting pin 22 high. This creates a pulse which triggers the A/D converter. Double buffering allows the UAR/T to be sending one word while the A/D converter is busy loading a new word. Thus the rate at which information can be sent is determined by the rate at which the UAR/T can send out the bits.

Sampling theorem states that at least two samples per second must be taken per hertz of data bandwidth. Thus, in order to represent an EKG which contains components up to 50 Hz, the analog EKG data must be converted to digital at least 100 times a second. Since

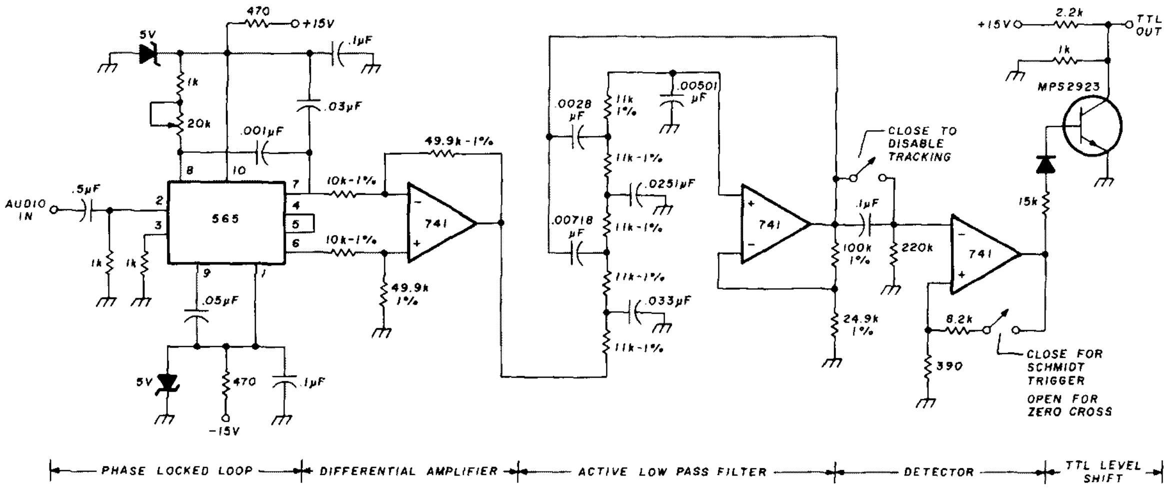



**fig. 7. The schematic diagram for the fsk demodulator. An active RC filter allows reception of the 1200-baud data.**

on a cassette tape to be transmitted later through the satellite. A block diagram of the system is shown in **fig. 4.** Each of the three lines takes place at a different time due to equipment locations and schedules.

The fsk generator is fairly straightforward and is shown in **fig. 5.** Basically an 8038 IC function generator is switched between two adjustable trimmer resistors giving independently adjustable *mark* and *space* frequencies. The output is phase coherent although switching does not necessarily take place at the zero crossing points of the sine wave.

## UAR/T description

As mentioned before, it is the UAR/T's job to perform the parallel-to-serial and the serial-to-parallel conversions. This IC does each job independently, simultaneously, and at different rates if desired. In my case this last capability was not necessary. As shown in **fig. 6**, the IC is programmed for an eight bit word, two stop bits, and even parity by setting pins 36-39 permanently high. The parity bit is included for error-detecting purposes and is used later to give an indication of the validity of the received data. The input data word is applied, in parallel, to pins 26 to 33 by the A/D converter and a pulse is generated by the flip-flops which loads this word into the UAR/T's internal register. The UAR/T begins each conversion produces an eight-bit word and the UAR/T adds four bits of bookkeeping data per word, we must transmit the bits at the rate of 1200 per second to process the EKG. It would be nice to go faster, but bandwidth requirements become more restrictive when it comes to the fsk; to send 100 data points per second, the UAR/T must be clocked at 16 times the baud rate (1200 baud). Thus a 19.2 kHz clock is required. This clock should be crystal derived and applied to both the transmit and receive halves of the chip (pins 40 and 17).

The receiving portion is equally as simple. The serial bit stream is applied to pin 20. The received word is presented on pins 5-12 and pin 19 goes high to signal its arrival. This control signal is also applied to the UAR/T through pin 18 to signal the removal of the data. When pin 18 is set low, it resets pin 19 low. However, this does not happen instantly and internal propagation delays turn the pin 19 signal into a short low-going strobe. If needed, this may be used to signal the arrival of a new data word.

The rest of the pins are control functions which should be wired as shown. Pin 21 must be strobed high when power is first applied to clear all internal registers and ready the chip for operation. Anyone desiring to build a modern RTTY system using surplus keyboards will find the UAR/T very handy indeed. Several manu-

facturers, including General Instruments, Texas Instruments, American Microsystems, and National Semiconductor, produce compatible ICs. The National MM5303N was used in the system I built.

## digital decoding

Receiving and decoding the fsk signal involves the same process as receiving RTTY. The only differences are that the system is running much faster and that the keying is a change in audio frequency presented to the transmitter rather than a change in its vfo. Fig. 7 shows the diagram of a suitable fsk demodulator. Note the lack of bandpass filters on the input. Since these tests were done on Oscar's experimental orbits, a clear channel was available. This made it easy to control Doppler shift. Once again, a phase-locked loop tracks the input signal frequency and provides an appropriate error signal. Following the differential amplifier, the error signal is applied to a five-pole, Butterworth lowpass filter which has a cutoff frequency at 1500 Hz. The output of the filter resembles a digital signal but shifts up and down from zero depending on Doppler and the receiver tuning. Therefore, a zero-crossing detector is useless at this point. The easiest way to remove the dc offset is by capacitor coupling. The time constant associated with this capacitor is chosen to block slow changes such as Doppler and slow receiver tuning changes, but to pass the faster incoming data. Now, you can use either a zero-crossing detector or preferably, a Schmidt trigger.

The next stage can serve as either type of detector. By adjusting the feedback resistance on the op amp, the hysteresis can be tailored to the amount desired. More resistance results in less hysteresis. This method of detection allows any combination of *mark-space* frequencies to be used. As long as these frequencies do not drift too near the passband of the lowpass filter, any shift that exceeds the effective modulation introduced by the noise will produce a waveform that the detector can convert into a usable digital signal. The fact that these frequencies may drift is of no consequence. If, for some reason, it is desirable to disable this tracking feature, simply short the coupling capacitor in the detector as shown.

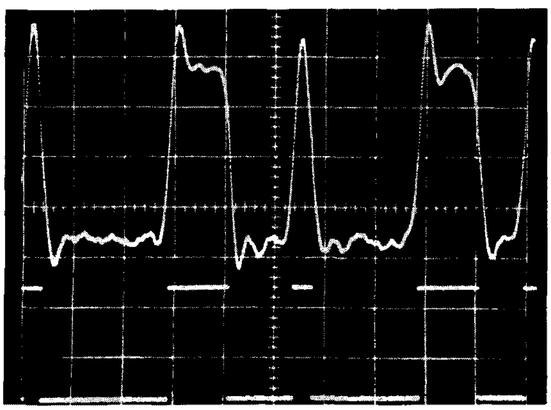


**fig. 8. Oscilloscope traces of the detector input (top) and the TTL output (bottom). The time base is 2 ms per division.**

This signal is next converted into a TTL compatible level. The net result is a decoder that does not care what the *mark-space* frequencies are, what their deviation is, and is not affected by drift, except by going too far down. On a test bench this circuit reliably decodes any shift above 50 Hz, with an input frequency of at least 1900 Hz and at any speed to 1200 baud.

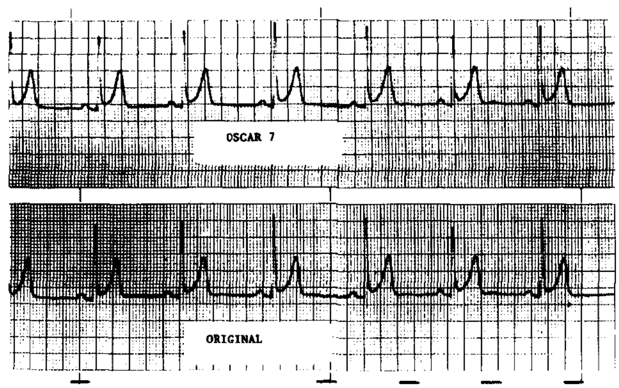



**fig. 9. Samples of the digitized EKG data as transmitted through Oscar 7 Mode B.**

Fig. 8 shows a dual-trace presentation of data as received through Oscar. The upper display is the input to the detector and the lower display is the digital output. The vertical scale represents shifts of 250 Hz per division; 850-Hz shift being used at this time. The waveform represents part of a digital word being received; note the noise-induced ripple. Any deviation exceeding this margin gives good data. This display shows how much deviation is needed for the noise level present. My first test used 170-Hz shift which proved to be insufficient as the noise margin was over 200 Hz. The next test, using 850-Hz shift succeeded. Subsequent tests are being conducted to determine the minimum necessary shift. Obviously the bandwidth presented to the phase-locked loop depends on the receiver and there is noise contribution from every bit of it. In my case the receiver was a Drake 2B with the selectivity set to 3.6 kHz.

Received signal strength also has a great deal of influence with respect to the noise. The use of bandpass filters on the input of the decoder could drastically cut this noise bandwidth but that would require the receiver operator to exactly track the Doppler shift by hand. In addition, the group-delay distortion of any such filters must be closely controlled since at 1200 baud you only have a couple of cycles of one frequency present assuming a 2400 Hz *mark*. To my way of thinking, it was the job of the decoder to eliminate the Doppler and not mine, so I just left the bandwidth wide open and used a wider shift. Only occasional retuning was necessary during the pass to compensate for the Doppler.

Following this decoder, the data goes to the receiving

half of the UAR/T where the serial-to-parallel transformation takes place. The parallel digital word presented al its output goes to a digital-to-analog converter which feeds a fourpole, lowpass filter set to 50 Hz. The output of this filter is the reconstructed analog data. Fig. 9 shows the result of the tests. The bottom strip is the original data and the top strip shows the same data after passing through Oscar 7 mode B. Out of 60 seconds of data, 5 parity errors were found. Although the same EKG was not used for both techniques, the improvement is obvious. Note the lack of degradation due to Doppler shift and noise. Doctors who examined the strips were unable to distinguish between the original and the satellite relayed versions.

While not optimum, this technique demonstrates what can be done with an ordinary audio channel. For more permanent experimentation I would recommend that you replace the 8038 fsk generator with a crystal-controlled programmable system, using digital counters, and synthesize a sine wave from a BCD-to-decimal decoder such as the 7442. This would give absolutely stable and accurate *mark-space* frequencies. Some input bandpass filters could be added to the demodulator. This system could also be adapted to multichannel use. It would require splitting the data word into two parts and sending each half, along with its appropriate address, as one word through the UAR/T. A substantially higher bit rate would then be required.

## summing up

So what has been the benefit of all this? Most obvious is that a satellite may be used to relay high-quality physiological data. This has definite use for regions where traditional land-based communications are marginal or non-existent. Small Alaskan towns are a prime example. Currently the federal government is involved in projects of exactly this nature. For amateurs this marks the first time that the FCC has granted permission to use an eight-level teleprinter code. At the time these experiments were being formulated, such permission did not exist and AMSAT was kind enough to pass my request along to the FCC. This opens the field for the use of surplus ASCII terminals and I anticipate that we will begin to see video terminals interconnected via amateur radio. As the cost of television typewriters continues to drop, this will make it easy for anyone to implement a low-cost RTTY station that can run at 25 characters per second. Although this experiment used an EKG, it could just as easily been text coded into ASCII and sent at 100 characters/second. The ability to handle bulky information rapidly can benefit users such as RTTY traffic nets, W1AW bulletins, and satellite telemetry of internal operations. The list of possibilities is limited only by one's imagination.

I need to acknowledge the help given me by Amsat in relaying my request for authority to use an eight-level code to the FCC, and to W6CG and W6ELT who helped control the satellite and transmit the tapes through it. I would be happy to answer anyone's questions upon receipt of a self-addressed, stamped envelope.

**ham radio**

# better audio

## for new or old communications receivers

Shaped audio responses can serve to reduce fatigue and distortion

A typical ssb signal can be sharp, penetrating, and with distortion painful to listen to. If your receiver's audio system has high distortion or lacks in volume, here is a technique that will solve the problems and open up a new concept in comfortable listening. It's also great for contests where distortion adds to the fatigue factor.

### the solution

Being a classical music fan and having a deep interest in audio, I came up with a technique that should attract any phone man. The audio system in many receivers leaves a lot to be desired. Exceptions to this were Karl Pierson and his KP-81 with high and lowpass audio filters. The Hallicrafters SX88 also had selectable audio response curves. Both had 10 watts of clean audio.

What I've done on all my receivers (except the KP-81 and SX88) is add a 20-watt hi-fi amplifier. Connected to the output of your receiver's detector (**fig. 1**), you can select either the normal receiver audio, the outboard amplifier, or both. Further, the use of a good speaker, preferably one of the good high-efficiency hi-fi speakers, means you don't have to increase the volume as much. A large speaker moves a lot of air and it "fills" the room. I often run upstairs and can hear the bigger speaker better upstairs as compared to a smaller one.

### audio emphasis

The ultimate in audio control is to use an audio amplifier with a Sound Effects Amplifier (SEA), as used by JVC in their audio systems. I find it very helpful in shaping the audio characteristics of various signals under different interference and voice conditions.

Radio Shack has an excellent SEA unit called a Stereo Frequency Equalizer that permits you to emphasize or attenuate any part of the audio response curve by 10 dB. The change in voice characteristics is useful under various receiving conditions and will also

By Ken Judge Glanzer, K7GCO, 202 South 124th Street, Seattle, Washington 98168

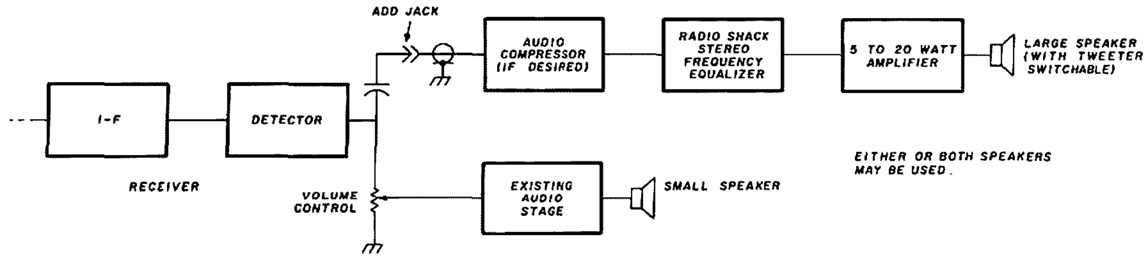
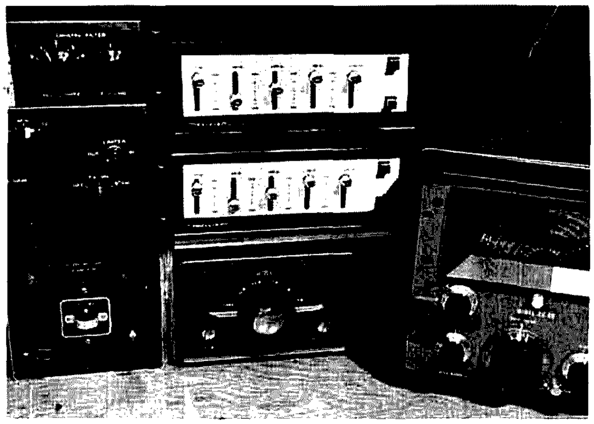



**fig. 1. The large speaker may be connected to the receiver through other pieces of audio processing equipment. This allows you to adjust for the best audio reception under various receiving conditions.**



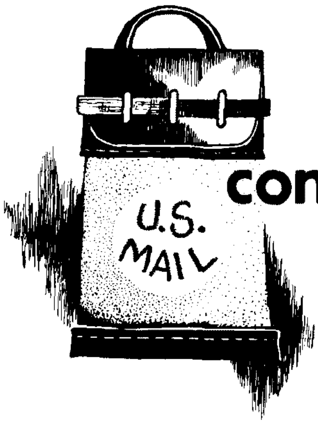
**The Stereo Frequency Equalizer can be used to shape the audio output of most receivers. They will provide attenuation or emphasis of specific frequency segments.**

act as an audio filter for CW. An equalizer could be used between the microphone and transmitter to provide the best audio for rag chewing or DXing. One idea is to record your normal and altered voice and then listen to the tape directly, not over the air.

## circuit modifications

Minor modifications can be made to many receivers by removing capacitors that are across the collector to base junction of transistors in the audio amplifier stage. In some cases there may be a capacitor across the primary of the output transformer. These capacitors usually attenuate the high frequencies. Typically the first loss in hearing occurs at the higher frequencies. This can be relieved by increasing the high frequency response. Sometimes the substitution of a different tube, transistor, or output transformer will also help to alleviate the problem.

**ham radio**




# comments

## RC notch filter

Dear HR:

A subscriber of your magazine recently showed me the article on "Hall's" notch filter in the September, 1975, issue* and I was very pleased with your enthusiastic presentation of the circuit. Obviously, I too feel that it has not had the popularity it deserves! Since you solicited comments on it and similar circuits, I would like to make a few comments which may be of some interest.

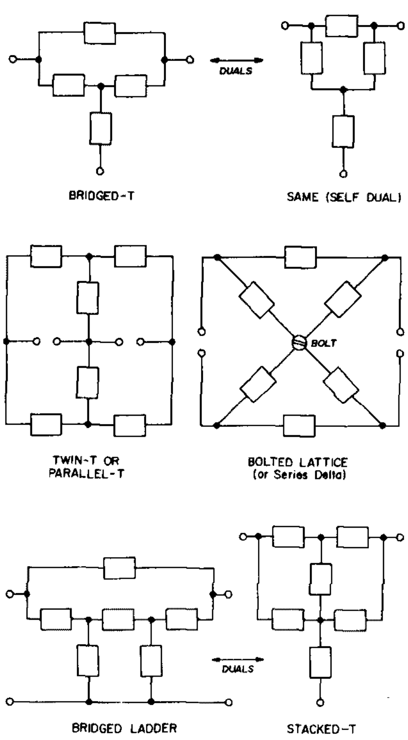



**fig. 1. Popular ac null circuits (left) and their duals (right). The bridged-ladder was the circuit discussed in WA5SNZ's article in the September, 1975 issue of the magazine.**

First, just for the record, the circuit has had some mention. The most complete description was by Leon Grillet *(91^c Congress des Societes Savantes*, Rennes, 1966, Tome II) which is in French, except for the algebra. You mentioned the article by Glasgal who referenced another *EEE* article by B.M. Van Emden (May, 1964). It has also appeared in a book, *Alternating Current Bridge Methods* (Hague and Foord, Pitman, London). This is Foord's revision of Hague's classic, the bible of ac bridge designers (my speciality). It was also briefly described by Penn and Grillet in letters (see *Electronic Engineering*, 1964). Others have mentioned it but only as introductory references before they described new circuits.

The name "bridged-differentiator" describes more than the network's typology which I've called a "bridged-ladder." It should be noted that the term "twin-T" has been used to describe measurement circuits with the same typology as the notch filter but with different circuit elements. Also, the dual† of a twin-T has been called a bolted-lattice (by an Englishman, I think) but could be called a "series-pi." The dual of the bridged-ladder might be called a "stacked–T" (see fig. 1).

The network has its advantages and disadvantages. The advantage in the General Radio null detector is not so much its economy on one potentiometer (which has a 46 dB exponential taper to get a log scale and is not inexpensive to make), but in its behavior. The obvious choice would have been a two-potentiometer Wien bridge, but when that circuit is adjusted in a high-gain feedback circuit, tracking unbalance as the wiper jumps from one wire to the next causes gain variations which make an indicating meter fluctuate widely. With one potentiometer, the gain may change in discrete steps as successive wires are contacted, but at least the peak is approached monotonically.

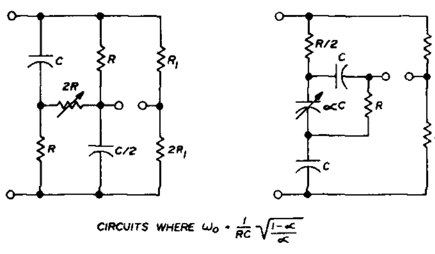



**fig. 2. Bridge circuits with a tuning law which tunes from zero to infinity. Over a 10:1 frequency range, a linear variable component provides a good logarithmic dial.**

The main disadvantage of this circuit is that its Q is lower than that of a twin-T, like all null circuits adjusted by one component, Q varies over the

*Courtney Hall, WA5SNZ, "Tunable RC Notch Filter," *ham radio*, September, 1975, page 16.

†Dual networks are defined as a pair of networks such that their branches can be marked in a one-to-one correspondence so that any mesh (or loop) of one corresponds to a cut-set of the other.

tuning range. The null detector uses the transfer impedance to obtain a more constant Q. The result is not ideal but surprisingly good. As I remember it, we were about to abandon the network when we discovered this way of using it.

There are other circuits to be considered, particularly the one by Y.A. Andreyev mentioned in the *G-R Experimenter* article (and by Grillett and Penn). Its tuning law seems preferable. There are many bridge-type networks which use a single variable resistor or capacitor. My favorites are those in the *G-R Experimenter* article that have a $\sqrt{(1-\alpha)/\alpha}$ tuning law which tunes from zero to infinity (see fig. 2). Over a 10 to 1 frequency range, a linear variable component gives a pretty good logarithmic dial. I've always wanted to experiment more with these circuits and maybe your readers would have fun making oscillators with them.

Finally, I would like to note that I'm very glad to have been made aware of your magazine While I'm not an amateur, the few issues of *ham radio* I glanced through contained several articles that interested me. I plan to look through more and follow it in the future.

**Henry P. Hall**
**Senior Principal Engineer**
**Gen Rad**

## 50-MHz frequency counter

Dear **HR**:

I am writing concerning WB2DFA's article on the "50-MHz Frequency Counter" in the January, 1976, issue of *ham radio*. The schematic for the counter circuit contains three errors in the crystal-oscillator section: the inputs to the SN7400 NAND gates are left floating. For the oscillator to work properly there must be a definite *high* on the unconnected input pins. Admittedly, the oscillator may work if the inputs drift high because of internal transistor action, but for reliable and stable operation there should be a *high* on the unconnected inputs to the NAND gates. This problem may be corrected in one of two ways:

1. Connect input pin 2 of U1A to pin 1, pin 4 of U1B to pin 5, and pin 10 of U1C to pin 9; or

2. Supply $+V_{cc}$ to the unconnected inputs through 2200- to 6800-ohm resistors (any resistor between these two values will work).

Both methods make the NAND gates function as inverters but method 2 decreases the effective fan-in of the chip so it presents less of a circuit load.

This should solve any problems readers may have with intermittent or malfunctioning oscillators in the frequency counter.

James R. Aiello
JRA Systems
St. Clair Shores, Michigan

*Although the inputs to the NAND gates should be pulled high rather than being allowed to float for true digital operation, we are not concerned with noise immunity in this case. In fact, it's less noise immunity that kicks the oscillator into action in the first place. Furthermore, tying both input pins together means that twice as much sinking current will be required to pull those inputs down, thus making the oscillator harder to start.*

*WB2WUF of Port Monmouth, New Jersey, built the counter almost a year ago, and in a recent conversation with him he reported that everything is still working okay. I have never had any problem with oscillator start-up either, and suggest that the circuit be left as is.*

James W. Pollock, WB2DFA

## telephone system precautions

Dear **HR**:

The telephone companies have rules against interconnects for a very good reason: they want the system to go on *working*.

In a recent issue of *ham radio** you published a schematic which showed a tap on a phone line which was directly grounded to the equipment. No doubt the author got away with it because he did not ground his equipment to the real world.

Although one wire of the phone pair into your house may appear to be at ground potential when first checked with a voltmeter, this is *not* a ground wire and the system must be isolated from ground to work properly. The phone will have all kinds of buzz and crosstalk if one lead is even bypassed to ground because phone pairs are *balanced* audio circuits.

A typical input to the switchgear in the phone office looks like this:

The transformer picks off the audio in a fully balanced manner, with the capacitor providing the tie at the center for audio. Another winding on the transformer passes the audio on into the system.

The relay detects the presence of dc current in the phone (off-hook and dial pulses), again in a balanced manner with two coils. Adding another ground at the user's end shorts one winding of each.

From there on, things get quite complicated because the relay is not always the same relay. A different physical relay may be in the circuit for *each digit* of the dialed number, with the dc provided during the talking phase being many steps removed (and perhaps many miles away) from the relay monitoring the line when on-hook. A ground on one lead causes great confusion in all this.

If you must attach to the system, there is a simple method of keeping out of trouble. *Always* present to the phone line a *transformer winding with a blocking capacitor in series*. The winding impedance can be anywhere from 50 to 2000 ohms. The capacitor should be no larger than 0.2 μF and able to stand 90 Vac ringing voltage with 48 Vdc superimposed (i.e. at least a 200 V paper capacitor).

On the other side of this transformer, you can do pretty much what you like as long as you don't drive audio into it at such levels that things are louder than normal speech in the normal telephone.

For a simple ring-detector, as was desired in the *ham radio* article, an optical coupler works admirably. Place a 0.1 μF in series with it to block dc from the phone system and put a diode across the optical coupler to pass the reversed part of the ringing cycle.

**N. J. Thompson, KH6FOX**
**Honolulu, Hawaii**

*Robert Shriner, WAØUZO, "Automatic Telephone Controller for Your Repeater," *ham radio*, November, 1974, page 44.

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# radio

## MAY 1977



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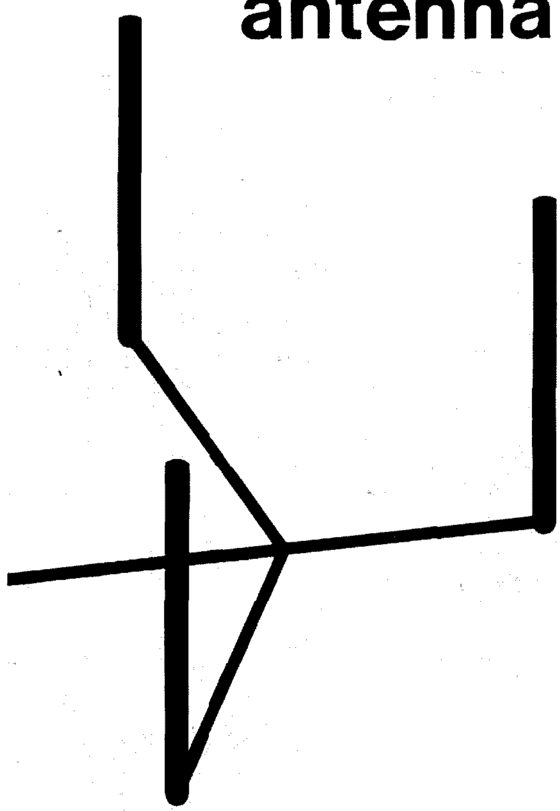

# ham radio magazine

## MAY 1977

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**James R. Fisk, W1HR**
**editor-in-chief**





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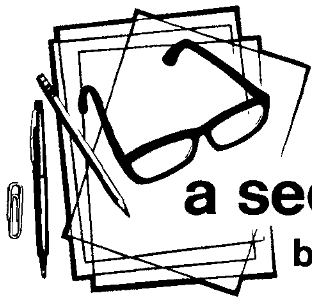

# a second look

by Jim Fisk

**The world distance record** on 1296 MHz was shattered on the 25th of January when two amateurs in Western Australia, Wally Green, VK6WG, in Albany, worked Reg Galle, VK5QR, near Adelaide, over a distance of 1886 kilometers (1171 miles). The previous record on 1296 MHz was set in October, 1973, between WA2LTM and W9WCD over a distance of 1240 kilometers (770 miles).

During their record-breaking contact VK6WG used CW and VK5QR was on ssb. Wally tried a-m, but Reg was only able to copy a few words because of the narrow passband of his receiver. There were participating observers at each end of the link, and much of the contact was tape recorded at VK6WG. It is more than likely that the two observers, Roger Bowman, VK5NY, and Bernie Gates, VK6KJ, were themselves green with envy – VK5NY, who was using a Microwave Modules MMc1296-LO converter, made a recording of the contact at his own station but was unable to get his signals through to VK6.

The transmitting lineup at VK6WG consists of an 8-MHz crystal oscillator/multiplier to 144 MHz, followed by an 832 tripler to 432 MHz, and a 3CX100A5 tripler to 1296. All equipment is homebrew. Power output on 1296 MHz is 10 to 15 watts; a pair of 807s are used as an a-m modulator. The antenna is a 3-foot (90cm) dish similar to that described in the RSGB's *VHF/UHF Manual*. The receiver front end was provided by Ron Wilkinson, VK3AKC; the converter was a Microwave Modules MMc1296 with a 28-MHz i-f.

The single-sideband gear used by VK5QR was an experimental hookup of a circuit suggested in 1970 by Karl Meinzer, DJ4ZC, in an article in *VHF Communications*. In this system the 432-MHz ssb signal is specially processed to eliminate most of the distortion caused by frequency tripling. The output from a homebrew 9-MHz ssb generator was mixed to 28 MHz, then fed to a 432-MHz transverter and 2C39A linear amplifier. The 432-MHz ssb signal was multiplied to 1296 MHz with a varactor tripler. Power output was about 10 watts to a 3-foot (90cm) dish. The 1296-MHz receiver consisted of a mixer-only converter into a low-noise preamp and tunable i-f.

This contact was a culmination of previous work by all four stations on 144 and 432 MHz. Both VK6WG and VK5QR deserve high praise for their stunning success, but in view of the relatively low power levels and small antennas used, and the good signal reports, there's a good chance the record may be bettered in the near future. There are a considerable number of amateurs east of Adelaide who are set up for operation on 1296 MHz.

**Jim Fisk, W1HR**
**editor-in-chief**

# presstop

FCC'S PROPOSALS TO BAN 10-meter amplifiers and require type acceptance of Amateur equipment (April ham radio) have drawn a letter from Senator Barry Goldwater to Commission Chairman Dick Wiley commending the Chairman on his recent pro-Amateur Radio statement, but questioning whether further restrictions on Amateurs are really going to solve the Commission's problems with illegal operators.

Type Acceptance would probably have little effect on illegal operators but would certainly raise equipment costs and delay product improvements for Amateurs, according to industry sources. In his statement that accompanied Docket 21116, the amplifier ban, Chairman Wiley expressed the hope that "the comments we receive will suggest other and better alternatives to the Commission's proposals." One such alternative is that already proposed to the Commission, by the San Antonio Repeater Organization, which would place legal responsibilities on the seller and buyer of transmitting equipment. The R.L. Drake Company is one enthusiastic backer of that approach and has already worked out a modification which would take the paperwork burden off the FCC but would still provide the Commission with traceability and accountability for enforcement purposes.

ELIMINATION OF AMATEUR SECONDARY STATION licenses proposed recently by the FCC would apply across the board — only individual primary station licenses would be retained. Existing secondary station licenses would remain in force until their expiration date, but could not be renewed. Comments on the proposal, Docket 21135, are due June 2; Reply Comments, June 30.

The Closed Season on issuance of new secondary station licenses applies only to applications received after March 3rd. All applications for renewal of existing secondary station licenses will still continue to be processed as before.

A 90-Day Delay of the May 25 due date for Comments on both Docket 21116 (Linear Amplifier Ban), and Docket 21117 (Type Acceptance of Amateur Equipment), has been requested by the ARRL because the two are "so far reaching in their implications for the Amateur Service that the Amateur community needs more time to study and respond to them." The League has also requested an extension on Comments due date for Docket 21135, the proposal to eliminate Amateur secondary station licenses.

HARMONIC AND SPURIOUS EMISSIONS from all Amateur transmitters will be specifically limited by a first Report and Order on Docket 20777, the bandwidth Docket. The new limits are 40-dB below carrier level under 30 MHz and 60-dB down between 30 and 235 MHz, and apply to all Amateur equipment — homebrew and modified as well as commercial.

The Proposal Is Causing Concern because it provides no relief for equipment already in Amateurs' hands so it could work a hardship on both the industry and individual Amateurs. Though add-on filters or antenna tuners could bring individual transmitters into compliance, as written, the Rules make no provision for external cleanup modifications.

AMATEURS MOVING to a new permanent location are now no longer required to advise the FCC of the change within four months as has been required by Part 97.95(a)(2) of the Rules. One caution must be observed however: be sure mail sent to you at the address the FCC does have gets to you, as "failure to reply to official communications" — a pink ticket for example — can get you in deep trouble!

"INSTANT UPGRADE" WAS AVAILABLE at all FCC Field Offices on March 1. The temporary authority for the successful applicant to use his new privileges is provided by a form filled out by the FCC examiner, and until the upgraded license arrives the Amateur must sign his call plus "Interim Washington" on phone (or "/WN" when on CW) if he took the upgrading exam in Washington — but only when he's exercising his newly won privileges. Each FCC Field Office has a 2-letter designator for use on CW.

FCC CHAIRMAN RICHARD WILEY was presented a plaque on behalf of the Amateur Radio Service and the members of the ARRL for "his excellent support of the Amateur Radio Service" by League President Harry Dannals at the Quarter Century Wireless Association's Washington Chapter banquet, which was attended by a number of top FCC and OTP officials.

AMSAT'S PHASE 3 SPACECRAFT FUNDING campaign is now officially in operation though a few details are still being worked out. It involves sponsorship of one or more solar cells from the spacecraft's solar panels at $10 per cell, a tax deductible donation. Send contributions to AMSAT Phase 3, Box 27, Washington, D.C. 20044 — sponsorship certificates will be sent to contributors.

AMATEUR RADIO INSURANCE plans are still being investigated by the ARRL with interesting proposals in hand. There's a good possibility that coverage in addition to simple protection of equipment will end up in the final package, which was referred to the League's Management Finance and Membership Affairs Committees for further study.

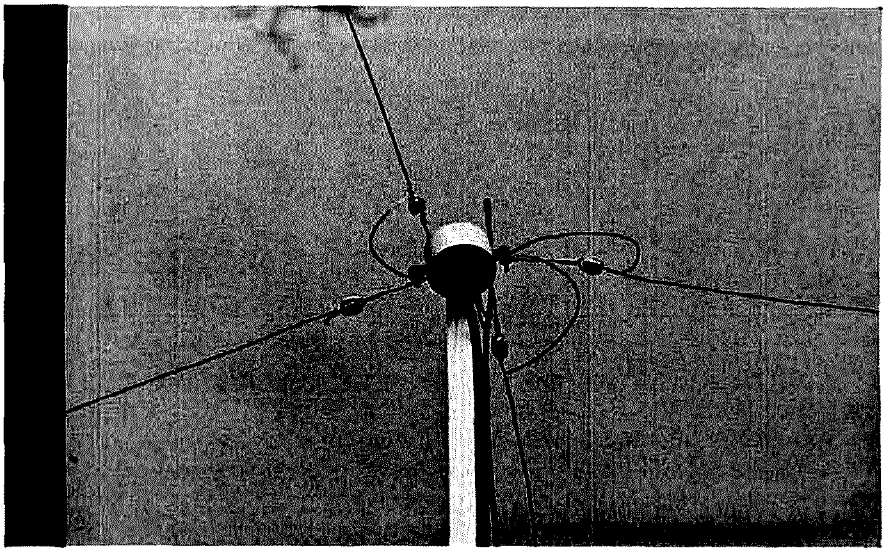

Photo by Douglas Fowley, WA4RPY

# antenna design

## using the longwire principle

A family of high-frequency amateur antennas with characteristics that should appeal to the experimenter and DX enthusiast

Amateur high-frequency antenna design has lagged behind the latest developments in amateur transmitters. The Ten-Tec and Heath solid-state transceivers, with their broadband output matching circuits, make transmitter adjustment as simple as tuning a receiver. The transceivers are easy to tune when terminated with a noninductive dummy load or a properly matched antenna operating near its resonant frequency. However, operating these transceivers on all bands becomes more complex. You must either switch antennas, or add an antenna tuner, or both. Broadband antenna designs, such as the discone and trap dipole are available, but each has its disadvantages. As a result of this restrictive situation, many amateurs operate near a frequency where the antenna resonates or operate in a narrow frequency band to avoid the nuisance of retuning.

### design features

This article presents an approach toward a new family of antennas using the longwire principle. Design features:

1. Outperforms conventional longwires of similar size during the majority of contracts under skip conditions.

2. Broadband characteristics from 80 through 10 meters are superior to any other antenna design.

3. Simple to feed using conventional 50-ohm coaxial cable.

4. No tuning or adjustment required if the design shown here is duplicated.

5. Design may be varied to suit your needs for desired bands, bandwidth, and array size. Adjustment is simple.

6. Neat in appearance and unobtrusive. Cost is below $100 for 2-kW PEP capability.

### performance under skip conditions

A justification of feature 1 is appropriate. Does the "best" technical design achieve the best performance during varying propagation conditions? Which antenna is best when several good alternative designs are being considered? I had a feeling that on-the-air antenna performance during skip conditions on the amateur bands was unpredictable. Conversations with other amateurs familiar with the use of different antennas

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confirmed this feeling. The evidence was strong that antenna characteristics based on theory and ground-wave measurements were *not* duplicated under skip propagation conditions.

I decided to make some experiments to test bandwidth improvement and the performance of alternative antenna design under skip conditions. Results were rewarding. These bandwidth experiments resulted in a new broadband antenna design. Although these tests were not under controlled laboratory conditions, I was nevertheless satisfied that the new antenna design gave outstanding performance. The purpose of this article is to acquaint you with the concepts, test results, and final design details so that you may understand the simple principles involved and, if you wish, design your own version.

## objectives

The first objective was to develop an antenna with low swr on all amateur bands from 80 through 10 meters. The swr was to be 2:1 or lower but in no case to exceed 3:1 from band edge to band edge.

The second objective was to develop an antenna design with outstanding performance when compared with other antennas of similar size, cost, and complexity. Emphasis was to be placed on simplicity and low cost. The performance judgement was to be based on the relative signal strengths achieved under medium and short skip conditions on 80 and 40 meters with secondary emphasis on long skip DX and higher frequency bands. The successful antenna would provide that big signal that all amateurs strive for.

## physical properties

Since I am blessed with a heavily wooded area and friendly neighbors, I decided to experiment with longwires. The operation was a low-budget affair from beginning to end. I used inexpensive aluminum electric-fence wire, which was purchased from Sears for about

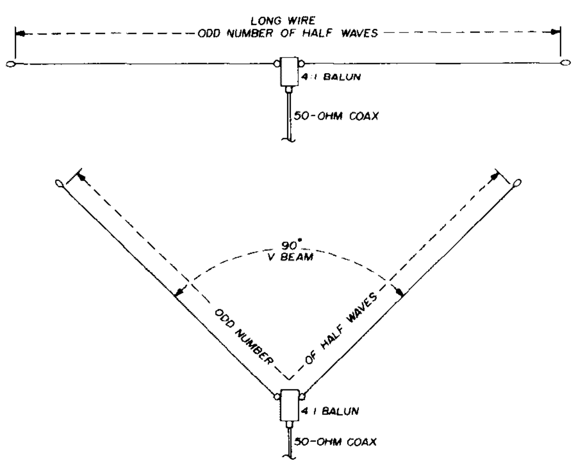



fig. 1. Basic elements author used to form 23 different antenna configurations for comparative signal-strength tests under skip conditions. Measured impedance at the feedpoint of both antennas was 200 ohms, which gave an excellent impedance match using the feed method shown.

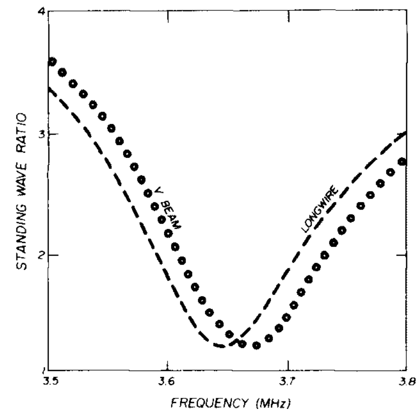



fig. 2. Swr measured on the two basic elements shown in fig. 1. At resonance the swr was 1.2:1; bandwidth at the 2:1-point was 120 kHz, or about 3% of the resonant frequency for both longwire and V beam.

\$15 for a quarter-mile roll. No insulators were used except at the ends of the wire. The wire was strung from tree-to-tree about 25 feet (7.6m) from the ground and laid in branches near the tree trunk.

To be acceptable to neighbors who owned the adjoining properties, the antenna had to be unobtrusive. The casual observer would have trouble seeing the thin aluminum wire in the trees.

During the experimental phase of the project the ends of the wires were brought into the operating room to facilitate both switching and measurements. Measurements and on-the-air comparisons were made over a two-year period before the final design evolved.

An unusually efficient ground system was needed to provide lightning protection, particularly during dry periods when ground conductivity dropped. The size of the array made it particularly vulnerable to electrical storms.

## electrical properties

The basic antenna element consisted of an odd number of half-wavelengths, fed in the center. In the early stages I determined that the feedpoint impedance was a nominal 200 ohms as measured on an impedance bridge — regardless of the number of odd half waves in a single element or the use of several elements. This greatly simplified impedance matching. A 4:1 balun, fed by 50-ohm RG-8/U coaxial cable provided excellent swr performance, as will be shown later.

According to the literature, a longwire radiator, if center-fed, must be fed at a current loop.[1] The use of an odd number of half-wavelengths at the resonant frequency results in conventional performance based on the entire wire length.

Where V-beam configurations were used, each leg was one-half the "element" length referred to above. Thus, each leg is one or more half wavelengths long, plus one quarter wavelength. This length results in the convenient impedance of 200 ohms at the apex. This technique of adding a quarter wavelength to the length of each leg to achieve impedance matching is one of the unconventional features of this design. The impedance of 200

ohms remained constant when either a longwire was center fed or a V-beam of the same total length was fed at the apex.

Although the configurations to be tested included V-beam elements, they were not chosen for their theoretical gain. On 80 meters, the legs were one and one-quarter wavelengths long, which would provide only about 3 dB gain over a dipole. The low antenna height made even this small amount of gain questionable during skip conditions. The V elements were to be tested for their ability to deliver a signal in any direction; high reliance was placed on the large capture area of the array.

Multiples of two basic elements were to be tested: the V beam mentioned above and a simple center-fed longwire. **Fig. 1** shows the elements. The apex angle of the V was chosen as 90° partly because this was optimum for whatever gain this element provided at an electrical length of one wavelength on 80 meters. On the higher-frequency bands, gain was of little consequence because of deviation from the optimum apex angle. The primary reason for the choice was that two center-fed longwires at 90° made a symmetrical array with all feedpoints at one place. Further, numerous alternative configurations could be selected at this central feedpoint.

**Fig. 2** shows the swr bandwidths for both the V-beam and longwires. Both antennas were just over 100 kHz wide at the 2:1 swr points on 80 meters which is far

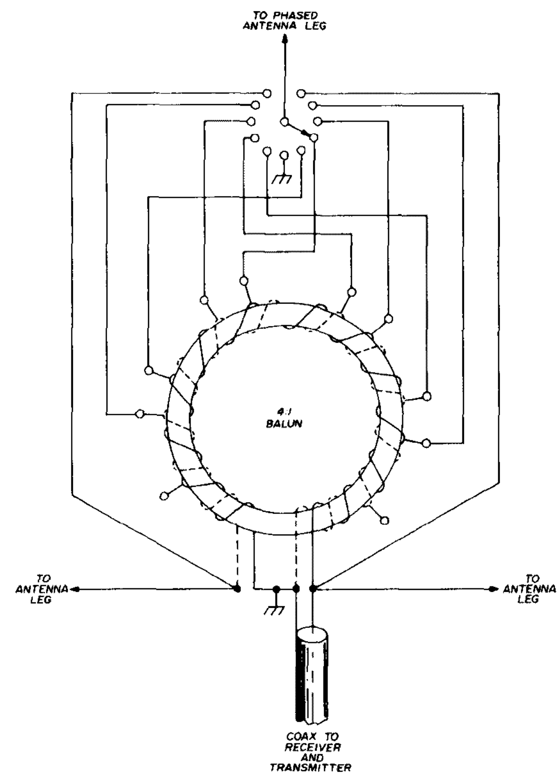



**fig. 3. Special tapped balun using Amidon Associates KW balun kit to test steerable phased arrays.**

from the desired 500-kHz bandwidth. To expand the bandwidth, I planned to stagger the element lengths.

One concern about staggering the element lengths was that such action would place the feedpoints off center in the element and adversely affect the impedance match. A check of this was hastily made, and it was a relief to find that the 200-ohm impedance measured by the bridge varied little with the imbalances contemplated. The swr measurements remained almost unchanged.

Terman[3] and others have described the steerable array. Changing directivity by changing phase relationships between multiple antennas was the usual method. I decided to attempt this by varying the phase relationships of several longwire elements and adding several steerable configurations to the on-the-air testing program.

Commercial steerable array designs were variable in the angle of radiation. They were most frequently used in a receiving system with phasing accomplished in the receiver circuits. It was my intention to test a high-level phasing system capable of transmitting as well as receiving. To test the steerable arrays, it was necessary to provide phase changes that could be switch-selected over a wide range. I made a special 4:1 balun with multiple taps as shown in **fig. 3.** I used a balun to test the effect on the transmitted and received signals by switching various phased elements during the on-the-air comparisons. I found that swr varied little when the phased arrays were switched.

## electrical environment

I gave some thought to the ground conductivity in the area to be occupied by the test antennas. Previously, using conventional antennas, I observed that performance deteriorated during the dry summer months. This seemed to indicate that performance was directly related to ground conductivity. A heavy rainy period in the summer months restored performance.

This variation in performance, apparently due to changed ground conductivity, raises the broader question of how the antennas under test would perform in other locations with changed ground characteristics. It seemed logical to assume that, if the tests were repeated during dry and wet periods, a measure of the validity of the tests in differing ground conditions would follow. Was it possible that an inferior antenna over poor ground would become superior over good ground conductivity? Would the ultimate be two antennas, one for good and one for poor conductivity? Fortunately, test results were consistent in both wet and dry conditions. No discernible difference could be identified. I therefore concluded that test results would be valid for other locations and that other amateurs would most probably experience similar results. Another constant was antenna impedance. The bridge and swr measurements varied little between wet and dry months.

Unfortunately, it was impractical to vary antenna height. The low height probably caused a high angle of radiation, which was satisfactory in medium and short skip, but would it work DX? My primary interest was in short-distance contacts with occasional DX work. I

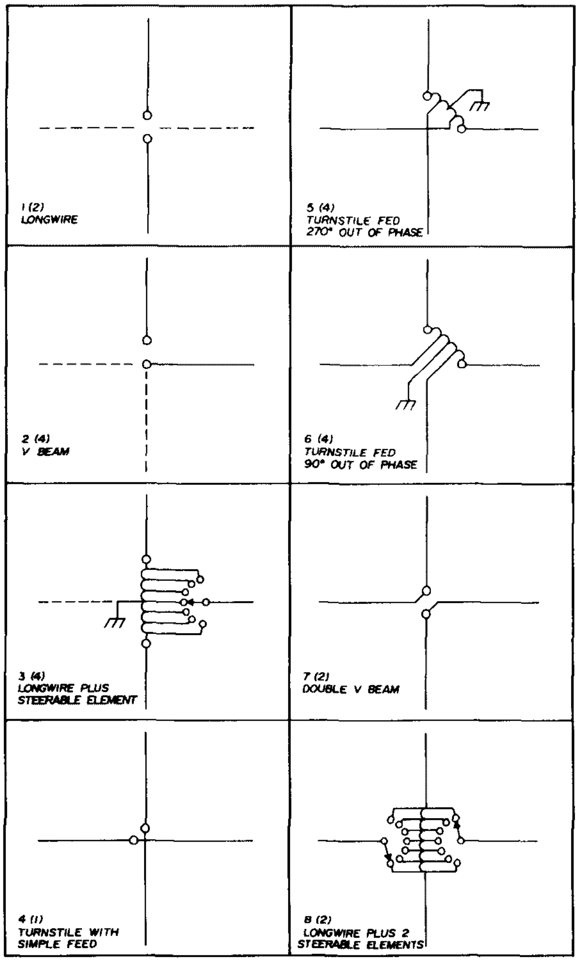



**fig. 4. Eight basic antenna designs tested under skip conditions. Numbers in parentheses signify use of additional legs for variations.**

decided to check low-angle radiation comparisons between the various antennas to be tested on a few DX contacts.

Nearby objects in the antenna site were relatively small and consisted of four houses, two power lines, and my electric fence. These objects undoubtedly had some effect on the antenna system, but it's unlikely they played a significant part in favoring one antenna configuration over another. As mentioned earlier, testing was not performed under laboratory conditions.

## testing

Most amateurs erect a single antenna for a specific band or bands. Rarely are alternative designs compared. It was a shock to me to observe that a very substantial difference exists between comparable antenna systems. If power is increased from 100 to 1,000 watts, gain is predictable (10dB). In a number of cases a difference of 30 - 40 dB was measured between antenna types. In one instance I tuned the 80-meter band during daytime and found it completely dead. Another antenna configuration was switched in and the band produced an S9 station about 300 miles away calling CQ. I made contact with S9 signals both ways. The original antenna was switched in on both transmit and receive several times with the same result – absolutely no signal. I tried a third antenna configuration, which also yielded absolutely no signal. Only the one antenna provided communications at S9 both ways.

The three antennas in the preceding instance were comparable in size, cost and complexity. They were various combinations of the same longwires. I concluded that, when an antenna is chosen for its theoretical performance in free space or over perfect ground, that antenna may be completely out of context with theory for producing a maximum signal during many propagation conditions.

Ideally, a number of antennas should be available and the best one selected based on its performance at the moment. This technique was possible and seriously considered, but I hoped a configuration could be found during the testing program that would produce the best or nearly best signals in the vast majority of cases. This would eliminate the nuisance of continually monitoring signal strengths on several antennas. This simplification was especially valued after I discovered that during some propagation conditions the optimum antennna shifted during a single contact.

Testing produced quite a few surprises, but the most valuable result was that one configuration proved to be superior. The need to switch several alternative antennas was greatly diminished. The superior configuration has never been covered in the literature to the best of my knowledge.

## test configurations

The basic test elements were center-fed longwires an odd multiple of a half wavelength long. Principal testing was on 80 meters, with fewer tests on 40 and a few

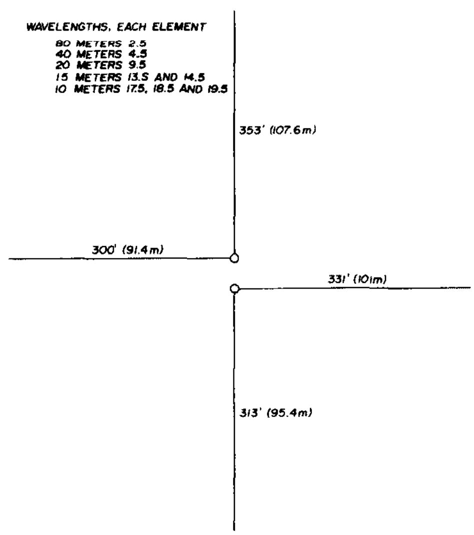



**fig. 5. Author's final design, the "double V."**

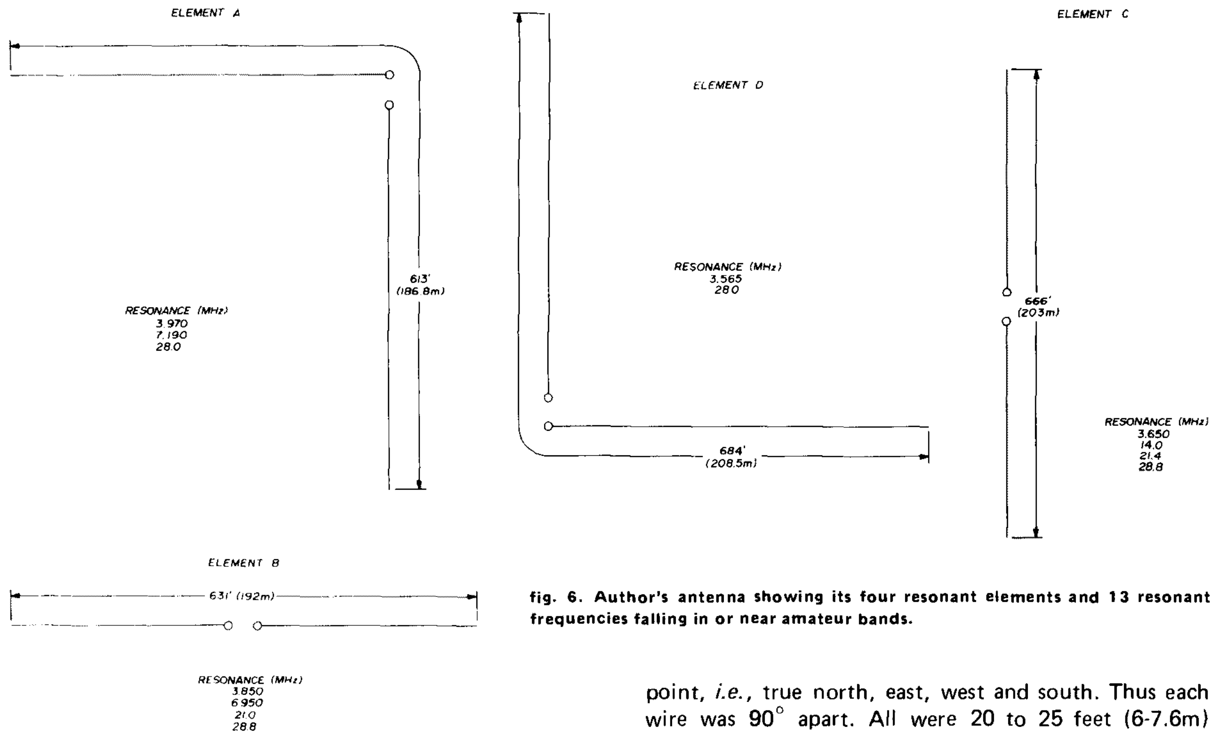



**fig. 6. Author's antenna showing its four resonant elements and 13 resonant frequencies falling in or near amateur bands.**

checks on 20. Note that a 625-foot (191m) wire is 2½ wavelengths long at 3.9 MHz and 4½ wavelengths long at about 7.05 MHz. This arrangement provides coverage of portions of both 80 and 40 meters. Another example of the physical length is a wire 660 feet (201m) long, which is 2½ wavelengths at 3.7 MHz and 9½ wavelengths at 14.1 MHz, giving coverage of portions of 80 and 20 meters.

The test elements were 680 feet (207m) long. They resonated just below 3.6 MHz; all 80-meter tests were made near this frequency. On 40 and 20 I had to tune out some reactance while using this wire length. Note also that the wire length referred to was the total length of the center-fed longwire, or the total length of both V-beam legs.

Emphasis was on the 80-meter-band tests because of the resonant condition. Fortunately results on the higher-frequency bands were similar to those on 80 despite the reactance offered by the test antennas on these bands.

The test configurations are shown in fig. 4. Eight basically different electrical designs were used with 23 variations. The large number of variations occurs because it was possible to combine different wires to obtain the same basic electrical design — but in a different direction. For example, antenna 1 has two variations: a north-south and east-west longwire. The number in parenthesis (**fig. 4**) signifies the number of possible variations. Let's take a look at the test configurations (**fig. 4**).

1. **Longwire**: All four legs were 340 feet (104m) long. Each wire was strung in the direction of a compass point, *i.e.*, true north, east, west and south. Thus each wire was 90° apart. All were 20 to 25 feet (6-7.6m) above ground.

This arrangement provided the alternative of a center-fed, north-south or east-west longwire. Center impedance measured 200 ohms on an impedance bridge at 3.6 MHz. The system was fed with a 50-ohm coax cable and a 4:1 balun. At 40 and 20 meters I added capacitance across the balun to compensate for the reactance. Unused elements were grounded at the operating position.

2. **V-beam**: Four V beams could be selected in different directions. Each leg was 1¼ wavelengths at 3.6 MHz.

3. **Longwire plus steerable element**: Four variations were possible with a switch-selected phase shifter connected to the steerable element. The phase shifter was a tapped 4:1 balun.

4. **Turnstile — simple feed**: This configutation is called a turnstile for want of a better name. The conventional turnstile has two half-wave elements spaced 90° around a common axis and is fed 90° out of phase. Before you dismiss this one on technical grounds, be advised that it performed very well.

5. **Turnstile 270° out of phase**: The phasing for this system was with a tapped 4:1 balun.

6. **Turnstile 90° out of phase**: Similar to no. 5 **but** with taps selected at 90°.

7. **Double V-beam**: Terminology is difficult for this one. It's not two V **beams** operating together in the usual arrangement with phasing lines. Again keep an open mind, because this antenna was a top performer.

8. **Longwire plus two "steerable" elements**: This

switched arrangement is really a combination of three antennas. At the extreme switch positions (as shown), the antenna is identical to no. 7 (double V-beam). At the center switch position the antenna is a simple longwire with the unused elements grounded. At the other intermediate positions, a variable phase shift is inserted between the longwire elements.

## test procedures

Initial tests were made on 80 meters with two-way contacts. Signal reports were obtained on the alternative antennas while transmitting. Only reports obtained from the distant operator using an S meter were recorded. After about 20 reports of this type I concluded that the reports correlated closely with readings I observed when receiving the distant station. I made further comparisons on received signals only, which speeded up the process and allowed a greater number of observations to be made in the time available.

All tests were made using no more than three antennas at a time. These antennas were instantaneously switched to obtain a reading. Several readings were taken on each occasion to overcome fading conditions and to ensure accuracy. One "control" antenna was carried over from one series of tests to the next. The readings were taken over paths between 200 and 1,000 miles (321.8 and 1609 km) with the majority in the 300-500 mile (482.7-804.5 km) range. Although only 50 watts output was used, the antenna system proved so effective that several contacts were made with Europe on 80 meters with comparative reports obtained on several of the alternative configurations.

On 40 and 20 meters, a few two-way contacts confirmed the same results as experienced on 80 meters: comparative readings were equivalent on both transmitting and receiving.

## test results

Five series of tests were performed. Results are as follows:

**first test series.** A comparison of antennas 1 and 2 (longwire and V-beam, fig. 4) was made. Results were surprising. The V beam outperformed the longwire by a substantial margin (10-20 dB) except on very rare occasions. By comparison the longwire was inferior, which seemed strange in view of its good reputation. Horizontal directional gain didn't appear to be a factor in achieving the superior performance from the V. In fact, a V aimed in the wrong direction frequently was *slightly* better than the "correct" V. The north-east path, which was over somewhat lower ground (not more than 8-10 feet [2.4-3m] elevation difference on flat ground), produced the best results in all directions – probably because of better ground conductivity.

**second test series.** These tests included the longwire plus steerable element (antenna 3 of fig. 4). At the extremes of switch travel a sort of lopsided V is in the circuit, while at the switch center position the antenna becomes a longwire with steerable element grounded. These tests again proved the superiority of the V configuration. Results of these tests ranged from inferior performance at the center switch position to best performance for the "lopsided V," with gradual improvement in between. The "steerable" element merely steered me back to the V for the best performance.

**Table 1. Feedpoint displacement from the center of the array.**

| | element | | | | | | | |
|---|---|---|---|---|---|---|---|---|
| | A | | B | | C | | D | |
| element length, feet (meters) | 613 | (186.8) | 631 | (192.3) | 666 | (203) | 684 | (208.50) |
| resonances (MHz) | 3.970 | | 3.850 | | 3.650 | | 3.565 | |
| | 7.190 | | 6.950 | | 14.0 | | 28.0 | |
| | 28.0 | | 21.0 | | 21.4 | | | |
| | | | 28.8 | | 28.8 | | | |
| actual feed displacement feet (meters) | 6.5 | (1.98) | 15.5 | (4.7) | 20 | (6.0) | 11 | (3.4) |
| worst possible displacements, | 60 | (18.0) | 62 | (18.9) | 64 | (19.5) | 66 | (20.0) |
| feet (meters) | 33 | (10.0) | 34 | (10.4) | 17 | (5.2) | 8 | (2.4) |
| | 8 | (2.4) | 12 | (3.7) | 12 | (3.7) | | |
| | | | 8 | (2.4) | 8 | (2.4) | | |
| ideal displacements, | 120 | (36.6) | 124 | (37.8) | 128 | (39.0) | 132 | (40.2) |
| feet (meters) | 66 | (20.0) | 67 | (20.4) | 34 | (10.3) | 16 | (4.9) |
| | 16 | (4.9) | 24 | (7.3) | 24 | (7.3) | | |
| | | | 16 | (4.9) | 16 | (4.9) | | |

**third test series.** In this series I compared the turnstile with simple feed, turnstile with 270° phasing, and turnstile with 90° phasing (antennas 4, 5 and 6 in fig. 4. Antenna 4 has a slight edge over the other two test configurations. It seemed to be a good all-around antenna.

**fourth test series.** Here I tested antennas 2, 4 and 7 (fig. 4); *i.e.*, V beam, turnstile with simple feed, and double V beam. As I appeared to have top performers at this point, I decided to test extensively on 40 as well as 80 meters and to check alternatives on transmit as well as on receive. Some testing was also done on 20 meters.

Again, twenty contacts were made on 80 meters, where a transmitting report was received from the

distant station. For simplicity, only the N-E, S-W Vs were used. The azimuth bearing, distance, time, band conditions and comparative dB readings were recorded. Results were:

| | turnstile | V beam | double V |
|---|---|---|---|
| equal or best signal | 6 reports | 9 reports | 11 reports |
| outperformed other antennas | 2 reports | 4 reports | 7 reports |

Careful examination of the test data disclosed no correlation between bearing or distance in providing the best result. Antenna 7 (double V) frequently outperformed the others by 5 - 20 dB. Antenna 2 (V beam) outperformed antenna 7 by only 2 - 5 dB, when it excelled. The turnstile (no. 4) excelled by 4 dB on skip once and by 4 dB on groundwave once. Antenna 7 not only produced the best signal most frequently but did so by the greatest margin.

Subsequently, a larger bank of data comprising one hundred and sixty two readings was obtained during varying conditions by recording the differences of received signals. The results:

| | turnstile | V beam | double V |
|---|---|---|---|
| equal or best signal | 26% of contacts | 38% of contacts | 36% of contacts |
| outperformed other antennas | 12% of contacts | 45% of contacts | 43% of contacts |

Again the phenomenon of greater average dB gain was observed for antenna 7 double V when it did excel; when it was inferior to the others, it was not far behind.

The several contacts with Europe on 80 meters indicated no discernible difference between antennas 4, 2 and 7. The differences seemed to disappear on long skip. On 40 meters antenna 7 was the undisputed leader:

| | turnstile | V beam | double V |
|---|---|---|---|
| outperformed other antennas | 13% of contacts | 10% of contacts | 77% of contacts |

On 20 meters antenna 7 did even better. It provided the best signal on more than 80% of the readings on short, medium and long skip.

**fifth test series.** The fifth and final test series was an anticlimax. It involved antenna 8 (longwire plus two steerable elements). Turning the switch brought signals from minimum in the center position to maximum in either end position with rarely a difference between the end positions. These end positions are identical to antenna 7 (double V) and tend to confirm its excellence as found in the preceding test series.

## test conclusions

The attempt to devise a steerable array was a dismal failure. There may be other ways to achieve an effective phase shift between elements, but the tapped balun is not the answer. I concluded that phase shifts were lost in the skip paths. The longwire is an inferior configuration. The same labor and materials can be used to construct a much more effective V or double V.

While the 80-meter performance difference between the V and double V is somewhat marginal, another factor heavily favors the double V. This is its ability to provide broadband performance.

This part of the article describes specific broadband designs. Performance figures and other data are provided so that you can tailor your own longwire array to suit your real estate and bandwidth requirements.

The V beam proved to be outstanding in an unusual configuration which I call a double V. The double V can be broadbanded by staggering its element lengths and easily fed with 50-ohm coax and a 4:1 balun. It will operate on all bands or on selected bands depending on the length and number of legs. At the sacrifice of bandwidth, three legs may be used instead of four.

## the antenna

My antenna is the shortest practical design for adequate bandwidth to cover all HF amateur bands (except 160 meters). It has four legs approximately 20-25 feet (6-7.6m) above ground, which terminate on a 40-foot (12m) mast. The legs vary in length from 300-353 feet (91-107.6m) and are spaced 90° apart. The wires which are uninsulated run through trees and are supported by branches.

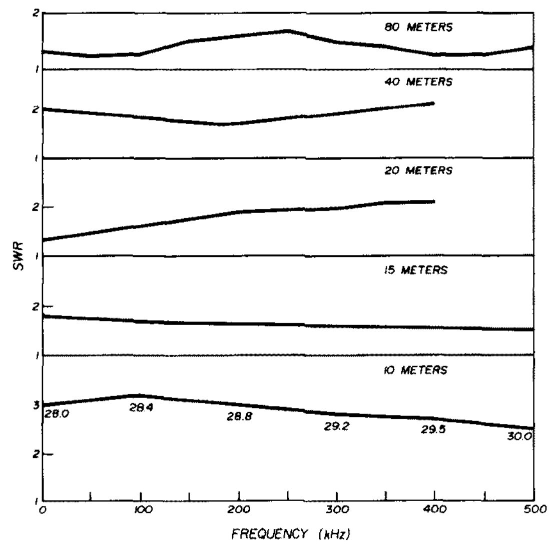
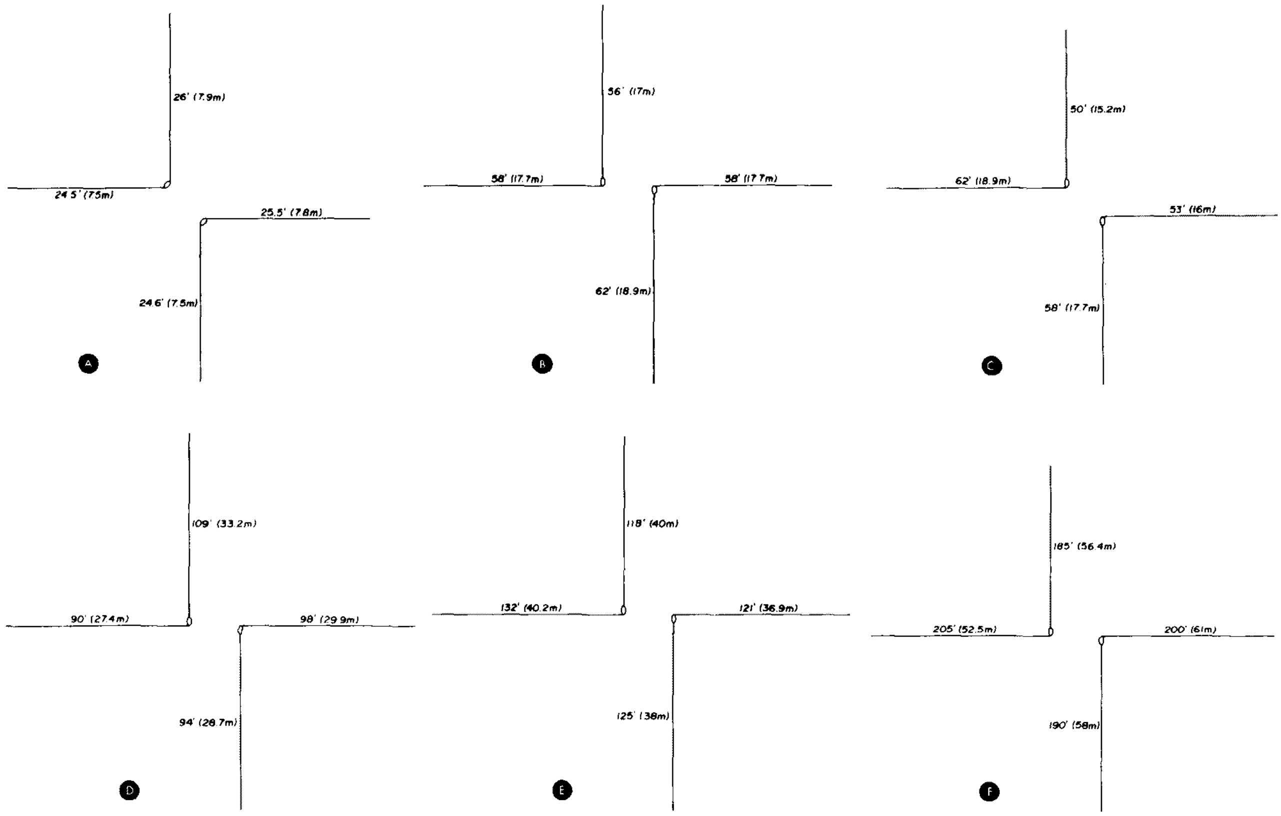



**fig. 7. Measured swr of author's antenna on all bands, 80-10 meters. The curve for 10 meters covers 28 to 30 MHz. The curves for the other bands start with the lower band edge on the left side of the chart.**

Aluminum electric fence wire is used, which can be purchased from Sears or Montgomery Ward for about $15 per quarter mile (0.4km) roll. The 5 kW PEP 4:1 balun by Palomar Engineers has stainless steel eyelets, which take 1/4 inch (M6) stainless nuts and bolts, and which I used for terminals. The terminals and wires were impregnated with zinc chromate metal primer paint to prevent corrosion.

Fig. 5 shows the final design. It's really a composite of four antennas with 13 resonances in the HF amateur bands as shown in **fig. 6**. The antenna is unobtrusive. Only a very alert observer would notice the small wire



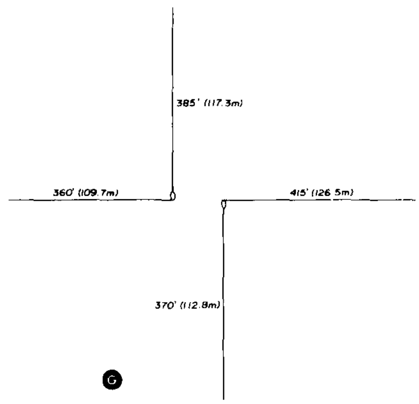




**fig. 8. (A) 10-meter broadband design. This is an extension of the simple 10-meter antenna referred to previously. Its four staggered leg lengths will provide a low swr across the entire 10-meter band. It should make a good OSCAR antenna for the high end of 10 meters as well as a good communications antenna for the low end of the band. (B) 15-and 10-meter "shorty" design. This antenna should do a good job across the entire 10- and 15-meter bands but with some sacrifice in bandwidth. (C) 20-, 15-, and 10-meter design. This arrangement would cover the low end of 15 and 20 meters and virtually the entire 10-meter band with low swr. Its performance would be best on 20 and 10 meters, where it has V beams, while resonating as a longwire on 15 meters. This is the smallest possible triband design. (D) 40-, 15-, and 10-meter design. This antenna would behave nicely on 40-meter CW, 15-meter phone and 10-meter phone. Operation on 40-meter phone would be satisfactory in most cases, but the swr would be borderline. This is the smallest possible design that includes the 40-meter band. (E) 20-, 15-, and 10-meter design. This is a beautiful broadband antenna for these bands with a low swr across the band in each case. The design provides V-beam resonances on each of the three bands for best performance. Place each leg at a compass point and it would be a satisfactory radiator in all directions. (F) 80-, 20-, 15-, and 10-meter design. This is the smallest possible design that includes 80-meters. It will give superb *performance* on that band. The swr will be below 2:1 across the entire 80- and 20-meter bands. On 15 meters the bandwidth will not be quite as broad, but the design should yield good results. The design resonates in V configurations on all four bands. The 10-meter-band swr bandwidth is somewhat unpredictable because of feedpoint displacements close to *current* nodes. (G) 160-, 40-, 20-, and 15-meter design. Antenna designers and manufacturers seem to studiously avoid provision for 160-meter operation. This is unfortunate in view of the interesting characteristics and challenges presented by this band. It is significant that some of the most sophisticated amateurs are now operating on 160 and that the LORAN system with its interference is all but obsolete. When the final demise of LORAN occurs, hopefully the 160-meter band will be restored to its pre-WW2 status. This antenna should be a winner on 160. It should also provide good performance on 40, 20 and 15 meters. On 10 meters, so many loops and nodes occur that performance is unpredictable, but you might be pleasantly surprised. This is the smallest design possible that includes 160 meters, and it's unfortunate that no resonances fall in the 80-meter band, which renders this band unusable except with a high swr. A combination of 160 and 80 meters in this design isn't possible unless wire lengths over 1000 feet (304.8m) are used — somewhat impractical for the average urban lot.**

snaking through the trees. Various parts of the system cross three adjacent properties with the permission of the owners.

With an array of this size, which covers some ten acres ($40470m^2$), lightning protection is essential. Secondary discharges from nearby strikes would produce lethal and damaging voltages. A direct strike would be devastating and would probably vaporize the wire. The objective should be to prevent these voltages from entering the operating position by providing a low

## 160 meters - 3% bandwidth: 60 kHz

| frequency (MHz): element length: full waves | half waves | 1.825 feet | (meters) | 1.875 feet | (meters) | 1.925 feet | (meters) | 1.975 feet | (meters) |
|---|---|---|---|---|---|---|---|---|---|
| 1.5 | 3.0 | 797.5 | (243.0) | 776.2 | (236.6) | 756.0 | (230.4) | 736.8 | (224.6) |
| 2.5 | 5.0 | 1338.1 | (407.9) | 1302.4 | (397.0) | 1268.4 | (386.6) | 1236.2 | (376.8) |
| 3.5 | 7.0 | 1878.9 | (572.6) | 1826.6 | (556.7) | 1781.0 | (542.8) | 1735.7 | (529.0) |

## 80 meters - 3% bandwidth: 120kHz

| frequency (MHz): element length: full waves | half waves | 3.6 feet | (meters) | 3.7 feet | (meters) | 3.8 feet | (meters) | 3.9 feet | (meters) |
|---|---|---|---|---|---|---|---|---|---|
| 1.5 | 3.0 | 403.2 | (122.9) | 392.3 | (119.6) | 382.0 | (116.4) | 372.2 | (113.4) |
| 2.5 | 5.0 | 676.5 | (206.2) | 658.2 | (200.6) | 640.9 | (195.3) | 624.5 | (190.3) |
| 3.5 | 7.0 | 949.8 | (289.4) | 924.2 | (281.7) | 899.8 | (274.3) | 876.8 | (267.2) |
| 4.5 | 9.0 | 1223.2 | (372.8) | 1190.1 | (362.7) | 1158.8 | (353.2) | 1130.0 | (344.4) |
| 5.5 | 11.0 | 1496.5 | (456.1) | 1456.0 | (443.8) | 1417.7 | (432.1) | 1381.4 | (421.0) |
| 6.5 | 13.0 | 1769.5 | (539.3) | 1722.0 | (524.9) | 1676.7 | (511.0) | 1633.7 | (498.0) |

## 40 meters - 3% bandwidth: 210 kHz

| frequency (MHz): element length: full waves | half waves | 7.0 feet | (meters) | 7.1 feet | (meters) | 7.2 feet | (meters) |
|---|---|---|---|---|---|---|---|
| 1.5 | 3.0 | 207.3 | (63.2) | 204.4 | (62.3) | 201.6 | (61.4) |
| 2.5 | 5.0 | 348.0 | (106.0) | 343.0 | (104.5) | 338.3 | (103.1) |
| 3.5 | 7.0 | 488.5 | (148.9) | 481.6 | (146.8) | 475.0 | (144.8) |
| 4.5 | 9.0 | 629.0 | (191.7) | 620.2 | (189.0) | 611.6 | (186.4) |
| 5.5 | 11.0 | 769.6 | (234.6) | 758.7 | (231.3) | 748.3 | (228.0) |
| 6.5 | 13.0 | 910.2 | (277.4) | 897.4 | (273.7) | 885.0 | (269.7) |
| 7.5 | 15.0 | 1050.3 | (320.1) | 1036.0 | (315.0) | 1021.6 | (311.4) |
| 8.5 | 17.0 | 1191.3 | (363.1) | 1174.6 | (358.0) | 1158.3 | (353.0) |
| 9.5 | 19.0 | 1332.0 | (406.0) | 1313.1 | (400.2) | 1295.0 | (394.7) |
| 10.5 | 21.0 | 1472.5 | (448.8) | 1451.8 | (442.5) | 1431.6 | (436.4) |
| 11.5 | 23.0 | 1613.0 | (491.6) | 1590.3 | (484.7) | 1568.3 | (478.0) |
| 12.5 | 25.0 | 1753.6 | (534.5) | 1729.0 | (526.5) | 1705.0 | (519.7) |

## 20 meters - 3% bandwidth: 420 kHz

| frequency (MHz): element length: full waves | half waves | 14.1 feet | (meters) | 14.2 feet | (meters) | 14.3 feet | (meters) |
|---|---|---|---|---|---|---|---|
| 1.5 | 3.0 | 103.0 | (31.4) | 102.2 | (31.2) | 101.5 | (30.9) |
| 2.5 | 5.0 | 172.7 | (52.6) | 171.5 | (52.3) | 170.3 | (51.9) |
| 3.5 | 7.0 | 242.5 | (73.9) | 240.8 | (73.4) | 239.1 | (72.9) |
| 4.5 | 9.0 | 312.3 | (95.2) | 310.1 | (94.5) | 308.0 | (93.9) |
| 5.5 | 11.0 | 382.1 | (116.5) | 379.4 | (115.6) | 376.7 | (114.8) |
| 6.5 | 13.0 | 451.9 | (137.7) | 448.7 | (136.8) | 445.6 | (135.8) |
| 7.5 | 15.0 | 521.7 | (159.0) | 518.0 | (157.9) | 514.4 | (156.8) |
| 8.5 | 17.0 | 591.5 | (180.3) | 587.3 | (179.0) | 583.2 | (164.0) |
| 9.5 | 19.0 | 661.2 | (201.5) | 656.5 | (200.1) | 652.0 | (198.7) |
| 10.5 | 21.0 | 731.0 | (222.8) | 725.9 | (221.3) | 720.8 | (220.0) |
| 11.5 | 23.0 | 800.8 | (244.1) | 795.1 | (242.3) | 789.6 | (241.0) |
| 12.5 | 25.0 | 870.6 | (265.4) | 864.5 | (263.5) | 858.4 | (261.6) |
| 13.5 | 27.0 | 940.4 | (286.6) | 933.8 | (284.6) | 927.2 | (282.6) |
| 14.5 | 29.0 | 1010.0 | (307.8) | 1003.0 | (305.7) | 996.0 | (303.6) |
| 15.5 | 31.0 | 1080.0 | (329.2) | 1072.4 | (326.9) | 1064.8 | (324.6) |
| 16.5 | 33.0 | 1149.7 | (350.4) | 1141.6 | (348.0) | 1133.7 | (345.6) |
| 17.5 | 35.0 | 1219.5 | (371.7) | 1211.0 | (369.1) | 1202.5 | (366.5) |
| 18.5 | 37.0 | 1289.3 | (393.0) | 1280.2 | (390.2) | 1271.2 | (387.5) |
| 19.5 | 39.0 | 1359.1 | (414.3) | 1349.5 | (411.3) | 1340.1 | (408.5) |
| 20.5 | 41.0 | 1428.9 | (436.1) | 1418.8 | (432.5) | 1408.9 | (429.4) |
| 21.5 | 43.0 | 1498.7 | (456.8) | 1488.1 | (453.6) | 1477.7 | (450.4) |
| 22.5 | 45.0 | 1568.5 | (478.0) | 1557.4 | (475.0) | 1546.5 | (471.4) |
| 23.5 | 47.0 | 1638.3 | (499.4) | 1626.7 | (495.8) | 1615.3 | (492.3) |
| 24.5 | 49.0 | 1708.0 | (520.6) | 1696.0 | (517.0) | 1684.0 | (513.3) |
| 25.5 | 51.0 | 1777.8 | (541.9) | 1765.3 | (538.0) | 1753.0 | (534.3) |

## 15 meters - 3% bandwidth: 630 kHz

| frequency (MHz): | | 21.1 | | 21.2 | | 21.3 | |
|---|---|---|---|---|---|---|---|
| element length: | | feet | (meters) | feet | (meters) | feet | (meters) |
| full waves | half waves | | | | | | |
| 1.5 | 3.0 | 68.8 | (21.0) | 68.5 | (20.9) | 68.1 | (20.8) |
| 2.5 | 5.0 | 115.4 | (35.2) | 114.9 | (35.0) | 114.3 | (34.8) |
| 3.5 | 7.0 | 162.6 | (49.6) | 161.3 | (49.2) | 160.5 | (48.9) |
| 4.5 | 9.0 | 208.7 | (63.6) | 207.7 | (63.3) | 206.7 | (63.0) |
| 5.5 | 11.0 | 255.3 | (77.8) | 254.1 | (77.4) | 253.0 | (77.1) |
| 6.5 | 13.0 | 302.0 | (92.1) | 300.5 | (91.6) | 299.1 | (91.2) |
| 7.5 | 15.0 | 348.6 | (106.3) | 347.0 | (105.8) | 345.3 | (105.2) |
| 8.5 | 17.0 | 395.2 | (120.5) | 393.4 | (120.0) | 391.5 | (119.3) |
| 9.5 | 19.0 | 442.0 | (134.7) | 439.8 | (134.1) | 437.7 | (133.4) |
| 10.5 | 21.0 | 488.5 | (149.9) | 486.2 | (148.2) | 484.0 | (147.5) |
| 11.5 | 23.0 | 535.1 | (163.1) | 532.6 | (162.3) | 530.1 | (161.6) |
| 12.5 | 25.0 | 581.8 | (177.3) | 579.0 | (176.4) | 576.3 | (175.7) |
| 13.5 | 27.0 | 628.4 | (191.5) | 625.4 | (190.6) | 622.5 | (189.7) |
| 14.5 | 29.0 | 675.0 | (205.7) | 671.9 | (204.8) | 668.7 | (203.8) |
| 15.5 | 31.0 | 721.7 | (220.0) | 718.3 | (219.0) | 715.0 | (218.0) |
| 16.5 | 33.0 | 768.3 | (234.2) | 764.7 | (233.1) | 761.1 | (232.0) |
| 17.5 | 35.0 | 815.0 | (248.4) | 811.1 | (247.2) | 807.3 | (246.1) |
| 18.5 | 37.0 | 861.6 | (262.6) | 857.5 | (261.4) | 853.5 | (260.3) |
| 19.5 | 39.0 | 908.2 | (276.8) | 904.0 | (275.5) | 899.7 | (274.2) |
| 20.5 | 41.0 | 954.8 | (291.0) | 950.4 | (289.7) | 945.9 | (288.3) |
| 21.5 | 43.0 | 1001.5 | (305.3) | 996.8 | (303.8) | 992.0 | (302.4) |
| 22.5 | 45.0 | 1048.1 | (319.5) | 1043.2 | (318.0) | 1038.3 | (316.5) |
| 23.5 | 47.0 | 1094.8 | (333.7) | 1089.6 | (332.1) | 1084.5 | (330.6) |
| 24.5 | 49.0 | 1141.4 | (347.9) | 1136.0 | (346.3) | 1130.7 | (344.6) |
| 25.5 | 51.0 | 1188.0 | (362.3) | 1182.4 | (360.4) | 1176.9 | (358.7) |

## 10 meters - 3% bandwidth: 840 kHz

| frequency (MHz): | | 28.2 | | 28.4 | | 28.6 | | 28.8 | |
|---|---|---|---|---|---|---|---|---|---|
| element length: | | feet | (meters) | feet | (meters) | feet | (meters) | feet | (meters) |
| full waves | half waves | | | | | | | | |
| 1.5 | 3.0 | 51.5 | (15.7) | 51.1 | (15.6) | 50.8 | (15.5) | 50.4 | (15.4) |
| 2.5 | 5.0 | 86.4 | (26.3) | 85.8 | (26.2) | 85.2 | (26.0) | 84.6 | (25.8) |
| 3.5 | 7.0 | 121.3 | (37.0) | 120.4 | (36.7) | 119.6 | (36.5) | 118.7 | (36.2) |
| 4.5 | 9.0 | 156.1 | (47.6) | 155.0 | (47.2) | 154.0 | (47.0) | 153.0 | (46.6) |
| 5.5 | 11.0 | 191.0 | (58.2) | 189.7 | (57.8) | 188.4 | (57.4) | 187.0 | (57.0) |
| 6.5 | 13.0 | 226.0 | (68.9) | 224.4 | (68.4) | 222.8 | (67.9) | 221.2 | (67.4) |
| 7.5 | 15.0 | 260.8 | (79.5) | 259.0 | (79.0) | 257.2 | (78.4) | 255.4 | (77.8) |
| 8.5 | 17.0 | 295.7 | (90.1) | 293.6 | (89.5) | 291.6 | (88.9) | 289.6 | (88.3) |
| 9.5 | 19.0 | 330.6 | (100.8) | 328.3 | (100.0) | 326.0 | (99.4) | 323.7 | (98.7) |
| 10.5 | 21.0 | 365.5 | (111.4) | 363.0 | (110.6) | 360.4 | (109.8) | 358.0 | (109.1) |
| 11.5 | 23.0 | 400.4 | (122.0) | 397.6 | (121.2) | 394.8 | (120.3) | 392.0 | (119.5) |
| 12.5 | 25.0 | 435.3 | (132.7) | 432.2 | (131.7) | 429.2 | (130.8) | 426.2 | (130.0) |
| 13.5 | 27.0 | 470.2 | (143.3) | 467.0 | (142.3) | 463.6 | (141.3) | 460.4 | (140.3) |
| 14.5 | 29.0 | 505.0 | (154.0) | 501.5 | (152.9) | 498.0 | (151.8) | 494.5 | (150.7) |
| 15.5 | 31.0 | 540.0 | (164.6) | 536.2 | (163.4) | 532.4 | (162.3) | 528.7 | (161.1) |
| 16.5 | 33.0 | 574.9 | (175.2) | 570.8 | (174.0) | 566.8 | (172.8) | 562.9 | (171.6) |
| 17.5 | 35.0 | 609.8 | (185.9) | 605.4 | (184.5) | 601.2 | (183.2) | 597.0 | (182.0) |
| 18.5 | 37.0 | 644.7 | (196.5) | 640.1 | (195.1) | 635.6 | (193.7) | 631.2 | (192.4) |
| 19.5 | 39.0 | 679.6 | (207.1) | 674.8 | (205.7) | 670.0 | (204.2) | 665.4 | (202.8) |
| 20.5 | 41.0 | 714.5 | (217.8) | 709.4 | (216.2) | 704.5 | (214.7) | 699.6 | (213.2) |
| 21.5 | 43.0 | 749.3 | (228.4) | 744.0 | (226.8) | 738.9 | (225.2) | 733.7 | (223.6) |
| 22.5 | 45.0 | 784.2 | (239.0) | 778.7 | (237.3) | 773.3 | (235.7) | 768.0 | (234.1) |
| 23.5 | 47.0 | 819.1 | (249.7) | 813.4 | (247.9) | 807.7 | (246.2) | 802.0 | (244.4) |
| 24.5 | 49.0 | 854.0 | (260.3) | 848.0 | (258.5) | 842.0 | (256.7) | 836.2 | (254.9) |
| 25.5 | 51.0 | 889.0 | (271.0) | 882.7 | (269.1) | 876.5 | (267.2) | 870.4 | (265.3) |

**table 2.** Center fed element lengths to produce nominal 200-ohm impedance at resonant frequencies shown. V-beam configurations are recommended using an apex angle of 90° with each leg one-half the wire length shown. Antenna will be virtually nondirectional and may be fed with a 4:1 balun and 50-ohm coax for low swr. Approximate bandwidths are shown for individual elements operating alone.

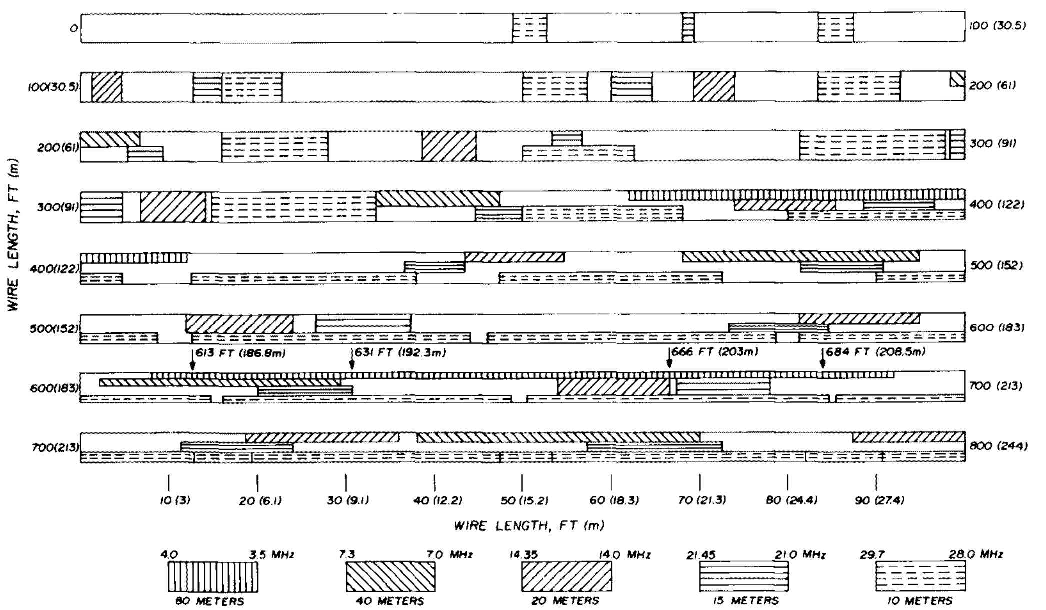



**fig. 9. Design aid showing wire lengths from zero to 800 feet (0-244m) in 100-foot (30.5m) increments with amateur bands identified for odd number of half-wave resonances. Wire length of author's antenna is identified by arrows.**

resistance path to ground. I installed a 9-foot-long (2.7m) ground rod at the base of the mast and ran a no. 4 AWG (5.2mm) copper wire from the balun to the rod. Thus, the entire system is grounded at all times.

**bandwidth**: One of the primary objectives was to achieve a low swr across all bands. **Fig. 7** shows the measured swr. It exceeded all expectations over the 80-meter band, where it was below 1.6:1. On 40, 20 and 15 it was below 2:1 from band edge to band edge.

The 3:1 swr on 10 was a surprise. From **fig. 6** you'll note that each of the four elements resonate on 10. The fact that they are fed offcenter because of the staggered leg lengths made calculating an impedance match difficult. This comes about because a mere 8-foot (2.4m) displacement from the element center produces a current node instead of the desired loop. Fortunately, impedance matching was quite tolerant so long as the antenna was not fed precisely at the node. This principle of considering the feedpoint displacement from the center in relation to the loops and nodes is quite important and is an inherent part of the design. The higher the frequency band the more critical this factor becomes. My antenna contained the following displacements as shown in **table** 1.

Examination of this data discloses an undesirable condition on 10 meters for elements A, C, and D. Only B is fed near the correct point. To correct this for 10 meters would result in bandwidth loss on the lower bands, so it was decided to accept this compromise. compromise.

Note that the relatively undesirable 15-meter-band feedpoint on element B may be balanced by the more desirable point on C, and the result is a tolerably low swr across the entire band.

The 20-meter performance is more difficult to understand. The only resonance within the band is given by C at 14.0 MHz. The feedpoint is displaced by 20 feet (6m) from center, which is quite close to the undesirable node. However, element D resonates at 13.7 MHz with a somewhat better match and this may account for the better swr skewed toward the lower end.

A probability that may explain the low swr on all bands is the combination of capacitive reactance of one element with the inductive reactance of another, which partially cancels the total reactance appearing at the feedpoint. A single resonant element alone appears to have a 2:1 swr bandwidth of about 3% of the frequency. However, it broadens out somewhat when other elements are added, even though they are not resonant near that frequency.

## design variations

An almost infinite number of variations of the basic design is possible. Three legs instead of four produce a combination of one longwire and one V. A three- or four-leg configuration may be designed for one or more bands with bandwidth also variable. The amateur who operates CW exclusively would favor the CW bands and design a smaller array to suit his purpose by sacrificing bandwidth to achieve smaller size.

To assist the reader in designing his own version, wire lengths are provided in **table 2**. Several frequencies in each band are given to make it convenient to stagger the wire lengths to achieve the desired bandwidth. Some pruning may be necessary after swr readings are made.

**Fig. 9** contains data for designing a multiple-band array. The wire is laid out in 100-foot (30.5m) increments, and the bands are identified where current loops occur in the center of the wire length. For example, a

50-foot (15.2m) length of wire will have a low swr on 10 meters if fed in the center with a 4:1 balun and 50-ohm coax. Its bandwidth will approximate 3% or 870 kHz. Thus, it will be useful from 28.5 to 29.4 MHz with an swr under 2:1. It should make a good antenna for 10-meter phone. If CW operation also is desired, let's add a leg 26.5 feet (8m) long to one of the balun terminals. This will give an additional resonance at about 28.2 MHz and result in a bandwidth from 28.0 to 29.5 MHz. Each active element will be 1.5 wavelengths long. As mentioned earlier, it's preferable to use the V configuration rather than the straight longwires. A number of alternate designs are shown in **fig.** 8.

## conclusion

The first advantage that impresses the operator who uses one of these broadband systems is the convenience. Gone is the need to adjust an antenna tuner. With a modern broadband solid-state transceiver, such as the Ten Tec or Heathkit designs, operation is instantly available in any portion of the band. Contest operating is greatly speeded up. When band conditions are marginal, band changing is immediate and the state of other bands can be checked with a flick of the band switch.

Operating habits will quickly change. Instead of concentrating in a small frequency range, the operator becomes a nomad who explores every nook and cranny of our broad frequency allocations. Interesting contacts are made with stations who rarely venture from their "home" frequencies.

The second advantage is performance. The capture area of the system gives a big signal on the band. When band conditions were poor and signals weak, I frequently experienced a pileup of weak stations calling in response to a CQ. An RST 599 report was common from stations flirting with the noise level.

Outstanding performance was not limited to the lower-frequency bands. The results in working long skip on 20 meter CW were the final pleasant surprise. Despite its low height, the antenna consistently turned in 589 and 599 reports from Europe and the Middle East during marginal conditions. ZL, VK and J stations could be heard and worked when other U.S. stations apparently could not hear them.

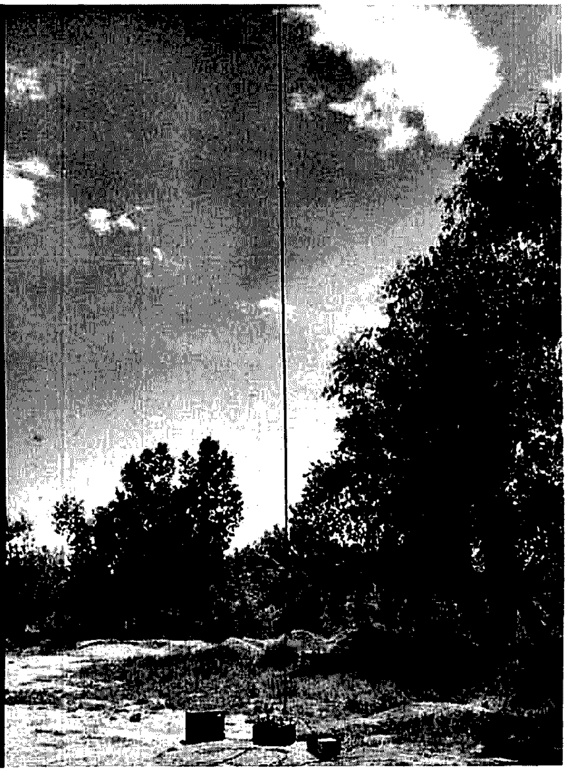
Operation on 10 and 15 meters was somewhat limited and I have yet to encounter the bands open for long skip. However, results were excellent on medium and short skip. Rarely did a station fail to respond to a call. Signal reports were outstanding.

In summary, the double V was the best all-around antenna I have used in some 40 years of hamming. While not tried as an inverted V, it would probably lend itself to such an arrangement when mounted on a tall tower.

## references


1. *The ARRL Antenna Handbook*, 13th edition, page 168.
2. F. E. Terman, *The Radio Engineer's Handbook*, 1st edition, 1943, McGraw-Hill, page 822.


**ham radio**



# ground screen — an alternative to a buried radial system

Using a ground screen to complement or replace the wire radial system to reduce ground losses

As an alternative to a radial system, the ground screen consists of a wire mesh of sufficient size to act as a capacitive connection to the earth, similar to a counterpoise, but laid on the ground instead of suspended overhead. It can consist of hardware or construction mesh, welded fencing, or even chicken wire, although the latter is less durable. I have been using a screen for fifteen years as a replacement for a radial ground system and find that it works quite well, although until recently I had been unable to make quantitative measurements on its efficiency.

A 1/2-wavelength wire radial system is a very effective ground return, although it has certain practical disadvantages. First, a lot of work is involved in installing the radials, especially if you move very often. Second, the area required for a low-band system is quite large and hurriedly-buried wires are prone to damage by the lawn mower. The ground screen has none of these drawbacks. My first ground screen was installed under a 60-foot (18.3m), base-insulated tower erected in my parent's back yard in suburban Cincinnati, Ohio. I tried this system since the driveways on each side of the tower left room for nothing else. The screen then consisted of two 15 by 5-foot (4.6x1.5m) lengths of construction mesh, one on each side of the tower.

I also tried a 1/4-wavelength flat top but found a single wire has such a small cross-section that it was an ineffective loading system for the tower. After many variations, I finally settled on a 65-foot (19.8m) sloping vertical wire with a 1/8-wavelength flat top. This system worked quite well both stateside and into Europe and was an effective testimony for the ground screen. Unfortunately, my beautiful base-insulated tower was a bust on 160 meters.

Several years later, I had my own home in Denver, Colorado, and could hardly wait to erect an extensive antenna system. Finances were short, however, so I settled on a 48-foot (14.6m) vertical made from 1-inch (25mm) aluminum tubing. A good radial ground system

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could not be installed under this vertical either, since the house and patio blocked the area to the South and the lot line was only 25 feet (7.6m) away to the North.

As before, I decided to install a ground screen due to the limited space available. I bought a 20 by 3-foot (6.1x0.9m) length of 1/2-inch (13mm) hardware mesh which was woven with no. 18 AWG (1mm) steel wire and then heavily dip galvanized. It was laid out along the house right under my 48-foot (14.6m) L-network-fed vertical, and the results were excellent on 40 and 75 meters.

**table 1. Ground screen configurations under the test vertical.**

| | |
|---|---|
| 1 | 10 x 3 feet (3x0.9m) with vertical centered |
| 2A | 20 x 3 feet (6x0.9m) with vertical centered |
| 2B | 20 x 3 feet (6x0.9m) with vertical 5 feet (1.5m) from one end |
| 3 | 30 x 2 feet (9x0.6m) with vertical centered (same area as no. 2) |
| 4 | 50 x 2 feet (15.2x0.6m) with vertical centered |
| 5A | 75 x 2 feet (22.9x0.6m) with vertical centered |
| 5B | 75 x 2 feet (22.9x0.6m) with vertical 25 feet (7.6m) from one end |
| 6 | 75 x 2 feet (22.9x0.6m) + 20 x 3 feet (6x0.9m) in a 90° cross with vertical centered |
| 7 | 45 x 2 feet (13.7x0.6m) + 20 x 3 feet (6x0.9m) in a 90° cross with vertical centered |
| 8 | 45 x 2 feet (13.7x0.6m) + 30 x 2 feet (9x0.6m) in a 90° cross with vertical centered (same area as no. 7) |
| 9 | 45 x 2 feet (13.7x0.6m) + 25 x 5 feet (7.6x1.5m) in a 90° cross with vertical centered |
| 10 | 45 x 2 feet (13.7x0.6m) + 30 x 2 feet (9x0.6m) + 20 x 3 feet (6x0.9m) in 60° radial strips with vertical centered |
| 11 | 45 x 2 feet (13.7x 0.6m) + 30 x 2 feet (9x0.6m) + 20 x 3 feet (6x0.9m) in overlapping parallel strips with vertical centered |
| 12 | 30 x 2 feet (9x0.6m) + 30 x 2 feet (9x0.6m) + 20 x 3 feet (6x0.9m) + 15 x 2 feet (4.6x0.6m) in overlapping parallel strips with vertical centered |

About a year later the vertical came down in an ice storm – I replaced it with a 50-foot (15.2m) tower and a four-element, 20-meter beam. For 40 meters I suspended a 45-foot (13.7m) vertical wire 5 feet (1.5m) from the side of the tower and tuned it with an L network, as I had done with the previous vertical. The tower affected the feed impedance of the wire, but did not seem to degrade its radiation since the wire vertical worked as well as a 40-meter inverted Vee.

A year and a half later the 40-meter wire vertical was replaced by a 40-meter beam on a 20-foot (6.1m) mast above the existing 20-meter antenna. With the additional top-loading provided by the 40-meter beam, it was now possible to use the existing 20 by 3 foot (6.1x0.9m) ground screen to shunt-feed the tower system on 160 meters. According to an article in *ham radio*,[1] the 65-foot (19.8m) tower with a two-thirds size 40-meter beam at the top and a four-element 20-meter beam 15 feet (4.6m) below should have an electrical length of about 110 feet (33.5m). The radiation resistance, however, is lower than that of a wire that long, so my ground screen was put to the test again. The tower ground-screen system was quite effective on 160 meters, and it allowed me to maintain weekly schedules through the 1976 season until May, over the 1100 mile path between Denver and Cincinnati. Reports up to 20 dB over S9 that were received from K1PBW made me decide that more quantitative data had to be obtained on the ground-screen idea.

In May, 1976, I went to WØSPM's farm with my GR916A rf bridge, GR1001-A signal generator, Drake R-4C receiver, three lengths of tubing for an antenna, and 210 square feet ($20m^2$) of new chicken wire for a ground screen. While chicken wire is no rival for mesh or fencing in a permanent system, it was much cheaper to cut it up into different lengths for experiments.

A 36-foot (11m) high, 1 inch (25mm) constant cross-section aluminum tubing vertical was erected and used for all of the following tests. This length was chosen since it was easy to install and could be part of an easily built phased array. Antenna impedance measurements were made on 1.8, 3.6, and 7.2 MHz. Various shapes and sizes of ground screen were tested under the vertical including a single strip, two lengths in a 90° cross, radial strips at 60°, and different overlapping lengths laid parallel to each other. The fourteen arrangements of screening are listed in **table 1**, while **table 2** gives the resistive and reactive antenna measurements for each case.

## resistive changes

Since no significant change in reactance was noted beyond 60 square feet ($5.6m^2$) of ground screen, only resistive changes will be discussed. These resistive values consist of the radiation resistance $R_r$ plus the ground loss $R_g$. When a single two-foot (0.6m) wide strip was used, the lowest total resistance ($R_r + R_g$) occured with a screen length of 50 feet (15.2m) on 1.8 MHz and 3.6 MHz, and 30 feet (9.1m) on 7.2 MHz. With greater lengths, the radiation resistance ($R_r$) probably increased as the screen departed from a lumped capacitance and began acting as part of a dipole leg. There was a similar occurrence with the long cross in **case 6** which resulted in a larger R value than with the smaller screen area of the **case 7** cross. **Case 9** yielded the lowest total R on all three bands.

**table 2. Impedance measurements for the 36 foot (11m) vertical with the different ground screen arrangements.**

| | 1.8 MHz | 3.6 MHz | 7.2 MHz |
|---|---|---|---|
| 1 | 25.0 – j730 | 28.0 – j286 | 85 + j67 |
| 2A | 18.7 – j739 | 22.0 – j285 | 80 + j72 |
| 2B | 17.7 – j739 | 23.0 – j278 | 80 + j71 |
| 3 | 13.5 – j728 | 18.5 – j281 | 78 + j81 |
| 4 | 11.0 – j719 | 20.0 – j274 | 81 + j81 |
| 5A | 12.8 – j711 | 21.7 – j276 | 82 + j71 |
| 5B | 12.6 – j717 | 21.5 – j275 | 81 + j81 |
| 6 | 11.2 – j714 | 18.0 – j281 | 74 + j77 |
| 7 | 8.8 – j717 | 16.6 – j278 | 74 + j76 |
| 8 | 8.8 – j719 | 15.7 – j279 | 72 + j78 |
| 9 | 7.9 – j717 | 14.6 – j279 | 71 + j76 |
| 10 | 8.2 – j717 | 15.6 – j279 | 72 + j76 |
| 11 | 8.2 – j717 | 14.5 – j279 | 73 + j78 |
| 12 | 8.5 – j717 | 14.8 – j279 | 72 + j76 |

The radial strip configuration of **case 10** had a resistive component slightly higher than in **case 9**, which was somewhat surprising. I thought it would have been equal to or better than that of the cross, but since more of the screen overlapped itself in this configuration, it is

probable that the decrease in area caused the increase in R. The ground resistance of the overlapping parallel strips was also slightly higher than the case 9 cross. I decided not to make the strips shorter and more numerous because of the difficulty in keeping all the pieces bonded together.

As would be expected, there were greater percentage changes on 1.8 and 3.6 MHz due to the ground losses ($R_g$) being a larger part of the R component. With approximately 200 square feet (18.6m$^2$) of screening, the cross (with one leg at least 5 feet [1.5m] wide)

A GR-916A rf-impedance bridge was used to make the impedance measurements. The chicken-wire mesh can be seen laid on top of the ground.

yielded the lowest ground loss. A single 30 by 3-foot (9.1x0.9m) strip of half that area, however, was only 38 per cent less efficient. This is the layout I am using under my tower at the present time.

## theoretical losses

Now, compare the theoretical impedance of a 36-foot (11m) vertical over a perfectly conducting ground with that of the test installation. This will enable the ground losses to be calculated. The following data, for 36-foot (11m) verticals, is from the *ARRL Antenna Book* graphs:

| | | |
|---|---|---|
| 7.2 MHz | 0.277 wavelength | 100° = 48 ohms |
| 3.6 MHz | 0.139 wavelength | 50° = 9 ohms |
| 1.8 MHz | 0.069 wavelength | 25° = 2 ohms |

Subtracting the theoretical from the measured values yields the following ground losses and efficiency

| | | |
|---|---|---|
| 7.2 MHz | 23.5 ohms | 67% |
| 3.6 MHz | 5.6 ohms | 62% |
| 1.8 MHz | 5.9 ohms | 25% |

The ground losses are higher on 7.2 MHz than were expected, but are probably due to the lower conductivity of the earth at this frequency. On the other hand, to obtain more than 60 per cent efficiency for a 1/8-wavelength vertical on 3.6 MHz is quite encouraging and might foster some interest in phased arrays on 80 and 75 meters, using only a ground screen instead of wire radials.

The 25 per cent efficiency on 160 meters is not spectacular, but compare it with the results of Brown, Lewis, and Epstein[2] using eight 135-foot (41m) radials on 3 MHz. With a 25 degree antenna they obtained only 27 per cent efficiency and a 5-ohm ground resistance, but a 50 degree antenna gave a 51 per cent efficiency with an 8.5-ohm ground resistance. The efficiency goes up because the radiation resistance increases 4.5 times, yet the ground resistance also goes up because the taller radiator induces currents into the earth at a greater distance from the antenna.

Using fifteen 135-foot (41m) radials on 3 MHz, Brown measured a ground resistance of 3.2 ohms. When he added a 9 by 9-foot (2.7x2.7m) copper screen the resistance dropped to 1.6 ohms. One hundred and thirteen radials 135 feet (41m) long (0.4 wavelength) brought this down to 1 ohm, but, unfortunately no information is available on the ground screen alone. One might use the parallel resistor equation however, to extrapolate from his data a probable ground screen resistance of 5 to 6 ohms. These values are very similar to the 5.6 and 5.9 ohm reading obtained on 3.6 and 1.8 MHz with my test system.

With ground losses as low as 6 ohms obtainable by using only 200 square feet (18.6m$^2$) of ground screen, efficient vertical radiators are within the reach of most amateurs. While good efficiency is obtained with a 36-foot (11m) radiator on 80 meters, some sort of top loading is advised on 160. A 2-ohm radiation resistance is just too low for good efficiency with most amateur ground systems.

I feel the ground screen is a good alternative to the wire radial ground system. I've used it for a long time and am glad to see test measurements that confirm its effectiveness.

An additional aspect of wire screening is that its application does not have to be limited to one type of installation. If you have the room and initiative to install a radial ground system, add a screen, too. If the antenna is approximately 1/4 wavelength or shorter, a screen will reduce the ground losses in the high-current zone at the base of the antenna. For your next vacation or Field Day project, a screen made of chicken wire offers the added benefit of being as easy to roll back up as it was to unroll. It is available almost anywhere, so you do not have to bring it with you. The next time you erect a new vertical, or plan to improve an existing one, consider a ground screen as either a substitute for radials, or as an adjunct to an existing radial ground system.

I wish to thank George Heidelman, K8RRH, and Dr. Mike Lee for their help in preparing this manuscript.

## references

1. John R. True, W4OQ, "How to Design Shunt-Feed Systems for Grounded Vertical Radiators," *ham radio*, May, 1975, page 34.

2. G.H. Brown, R.F. Lewis, and J. Epstein, "Ground Systems as a Factor in Antenna Efficiency," *Proceedings of the Institute of Radio Engineers*, June, 1937.

**ham radio**

# a new coaxial balun

## Re-examining the basic balun and applying new techniques makes it a useful, wide-range tool

The basic and sole purpose of a balun is to prevent the coaxial feedline from radiating and becoming part of the actual antenna structure. This should be understood before its operation is explained. By connecting one side of the dipole to the outer conductor of the coax, the radiating structure shown in fig. 1 is formed. The coaxial cable connected to one side of the dipole and extending at right angles has currents flowing on the outside surface of the braid and will radiate just as effectively as the conductors making up the actual dipole.

Fig. 2 shows the effect using a balancing device. Electrically the dipole is now fed by the two generators. The important point to be understood is that the electric field in space is zero in the plane perpendicular to and passing through the center of the dipole. Because of this, conductors may be placed in this plane without changing antenna operation in any way. Thus, if the coaxial feedline is brought away from the antenna at right angles it will not disturb antenna operation; there will be no rf current flowing on the outside braid so there is no radiation from the feedline. Other conductors that are normally in this plane, such as a mast or the metal boom of a Yagi array, will also have no effect on antenna operation.

### typical baluns

The sleeve and half-wave loop baluns are two well-known balancing devices (figs. 3 and 4). The transformation ratio of the sleeve balun is variable by changing the characteristic impedance of the inner line. However, its chief disadvantage is the difficulty of construction. If flexible cable is used for the inner line, then connection of the sleeve section and support of the cable in the sleeve present problems. If the jacket is left on the cable, the correct balun length will be less than a free-space quarter wavelength and will depend on the diameter of the sleeve. Also, when used at the higher uhf frequencies, 432 MHz and up, the sleeve diameter must be kept small enough to avoid having a structure that supports higher order modes of propagation.

The half-wave loop balun of fig. 4 is simple to build using flexible coax but has limited usefulness since the

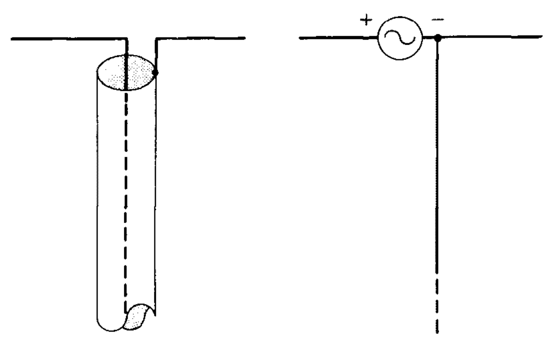


fig. 1. A dipole antenna fed directly with coaxial cable, left, and its electrical equivalent, right.

transformation ratio is fixed at 4:1. The characteristic impedance of the half-wave section, $Z_1$, has no effect on the transformation ratio or the balancing action of the device. The only effect for different values of $Z_1$ is to change the swr in the half-wave section, which is $2Z_0/Z_1$ when the load impedance, $Z_L$, is $4Z_0$. Typically, this balun is operated with $Z_1 = Z_0 = 50\ ohms$ and $Z_L = 200$

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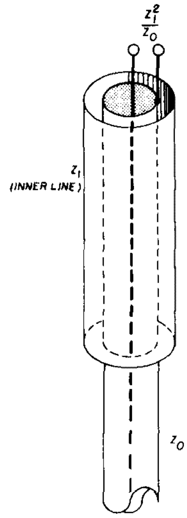


**fig. 3. The sleeve or bazooka balun has a fixed transformation ratio of 1:1. Its construction can be complicated by the lack of readily available materials for different impedance ratios.**

*ohms*. In this case there will be an swr of 2:1 in the half-wave section. To achieve unity swr in the half-wave section requires $Z_1 = 2Z_0$ which means a 100-ohm section if the main line is 50 ohms.

## a different balun

A coaxial balun configuration which is apparently new to amateur use is shown in **fig. 5**. The transformation ratio is adjusted by changing the characteristic impedance, $Z_1$, of the quarter-wavelength and three-quarter-wavelength sections. It will match a coaxial line (impedance $Z_0$) to a balanced load with an impedance of $Z_1{}^2/Z_0$. With this $Z_L$, the swr in the two short sections is $2Z_0/Z_1$ (or $Z_1/2Z_0$ if it is greater than unity), as in the half-wave balun. However, in this case the swr is set by the transformation ratio and will be unity only when $Z_L = 4Z_0$. You can see that the balun is simply a combination quarter-wave matching section and half-wave balun. The same thing may be achieved by using a regular half-wave loop balun and a quarter-wave matching section of either coaxial line or balanced line on one end. The variable configuration is useful because in a particular situation, the required coaxial impedance values may be easier to obtain.

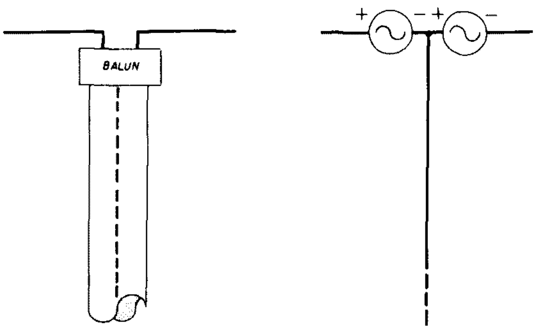


**fig. 2. A dipole fed through a balancing device will appear as two balanced generators. In this case there is no radiation from the feedline.**

As an example, consider the case where the main feedline has a characteristic impedance, $Z_0$, of 50 ohms. If the quarter-wavelength and three-quarter-wavelength sections are 50 ohms then $Z_1$ equals 50 ohms and you can now match a balanced load impedance of 50 ohms. The swr is 2 on the balun sections. Thus, a 1:1 transformation ratio is achieved using the same type coax in the balun as in the main feedline. It also avoids the construction difficulties of the sleeve balun by requiring only a coaxial T-fitting for connection between the coaxial line sections. Finally, note that multiband operation is possible with this balun as well as with the half-wave loop and bazooka baluns. For example, if cut for 144 MHz, they operate identically on 432 MHz and 1296 MHz. This assumes that the dielectric constant, and hence

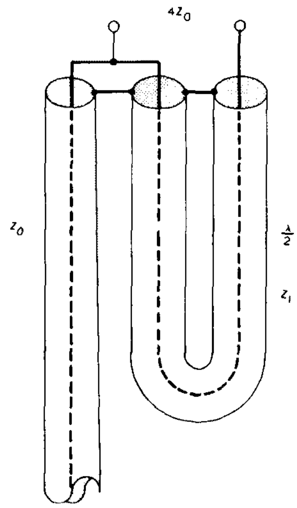


**fig. 4. The halfwave-loop balun has a normal transformation ratio of 4:1. The characteristic impedance of the matching section has no effect on the transformation ratio.**

velocity factor, in the balun sections is not appreciably different at the higher frequencies.

## multi-element coaxial lines

The variable transformer balun can be applied to a practical problem – the feeding of a high-gain Yagi antenna which presents a balanced load of 10 to 15 ohms to the transmission line. If the main feedline is 50 ohms and the load is 12.5 ohms, then $Z_0 = 50\ ohms$, $Z_L = 12.5\ ohms$, and the characteristic impedance of the balun sections is $Z_1 = \sqrt{Z_0 \cdot Z_L}$ or 25 ohms.

The problem of building coaxial line with low impedance is easily solved by using the multi-element coaxial line shown in **fig. 6**. Here, two 50-ohm lines are connected in parallel to obtain a 25-ohm line. The attenuation of the multi-element line is equal to the loss for the type of coax used. In general, any number of different types of coax cable may be paralleled as long as their velocity factors are equal. The result is a coaxial line with a characteristic impedance equal to the parallel combination of individual cable impedances. For

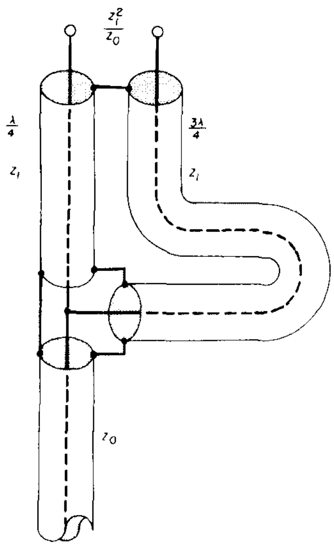


**fig. 5. The variable transformer balun as described by the author. By selecting the correct value of characteristic impedance for the matching section, the transformation ratio can be changed.**

example, 75-ohm and 50-ohm sections placed in parallel form a 30-ohm line. The sections can be soldered together or coaxial T-connectors may be used.

The solution to our matching problem, therefore, is shown in **fig. 7**. A system has been constructed that matches a balanced 12.5-ohm load and is built entirely of 50-ohm coaxial cable.

## shielded balanced line

Coaxial transmission line may also be used to form a balanced line when connected as shown in fig. 8. In this configuration, the characteristic impedance of the coaxial lines must be identical to maintain balanced operation. Also, the outer conductors must be connected at each end. The characteristic impedance of the balanced line is then twice that of the coax used. Two points should be noted. First, since the line is shielded, it may be run anywhere and, second, relatively low values of characteristic impedance may be obtained. A low impedance may be the simplest solution to a particular matching problem.

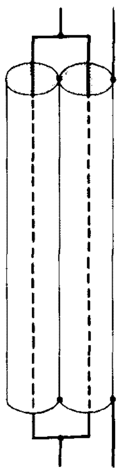

**fig. 6. A two-element coaxial section. The characteristic impedance of the composite line is the parallel combination of the individual sections.**

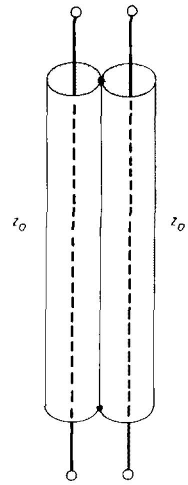


**fig. 8. Low impedance, balanced transmission line can be built by using two sections of coaxial cable. This line will retain the shielded characteristics of coaxial cable.**

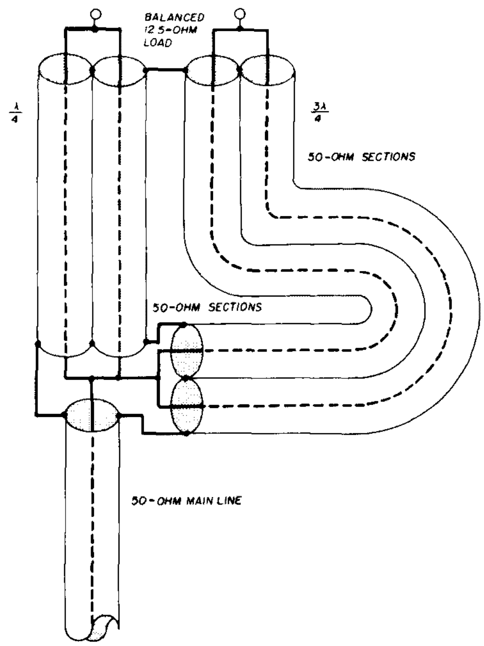


**fig. 7. A balun can be built by combining the techniques of composite lines with the variable transformer. This balun will match a 4:1 unbalanced-to-balanced, impedance ratio using only 50-ohm cable.**

## summary

I have tried to create a better understanding of the balun by discussing its purpose in an antenna system. A new configuration of coaxial balun has been presented, along with unusual coaxial cable techniques, that I believe will be found useful. At WAØRDX, the 50-ohm, 1:1 transformer balun, in particular, has proved to be a valuable alternative to the half-wave loop balun and bazooka balun for feeding vhf antennas.

**ham radio**

# antenna-transmission line analog

## a key to designing and understanding antennas

Practical applications of this valuable tool in designing your own antenna systems

In the first part of this article it was shown in some detail how an antenna may be regarded as possessing both radiating and non-radiating properties.[1] I also discussed how the non-radiating TEM wave mode function may be used to convert the antenna into a special kind of *rf* transmission line. Using the *antenna/transmission* line concept in a way first made clear by Dr. Schelkunoff,[2] you can determine the mean or average characteristic impedance *K* of this antenna/transmission line and use it to calculate the antenna's input impedance behavior either at a single frequency, or over an entire band of frequencies.

In this section I will show you how to apply the antenna/transmission line analog concept to a number of typical antenna design problems which arise in both amateur and commercial radio communications. No higher mathematics is needed to use the analog key in the modified form presented here. **Table 1** references the basic equations used, based on everyday trigonometric functions. However, to remove any possible difficulty for the interested amateur who may be a bit rusty in ac math, not only are all examples fully worked out, step-by-step, but the Smith chart[3,4] is also used to clearly present the progression of events leading to each solution of the antenna design problem.

The best way to understand something new is to plunge in and start using it. Let's begin with an antenna which is of increasing interest to the amateur who is faced with shrinking backyard space or a nasty tempered landlord: the electrically small antenna. Many forms of this antenna have been around for a long time, yet it is a tricky little beast and requires surprising care in its design if you want to obtain a reasonable level of performance from it.

### inductance loaded, electrically small antennas

When the electrical length, $2h_t^\circ$, of a doublet antenna is less than 180 degrees ($\lambda/2$), or the electrical length, $h_t^\circ$, of a grounded monopole is less than 90 degrees ($\lambda/4$) at the operating frequency, the antenna is too

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short electrically to oscillate in a state of natural resonance. For a linear antenna to oscillate at its first (lowest) natural resonance, the series inductance and shunt capacitance *distributed per unit length* along its conductor (or conductors) must add up to the reactive sum $+jX_L + (-jX_C) = jo\ ohms$ in a way similar to that of lumped, series LC circuits.

It is easy to understand how the distributed series inductance comes about, even in a perfectly straight conductor, but the distributed shunt capacitance is less obvious. The shunt capacitance is distributed to ground from each tiny incremental length of conductor forming a grounded monopole, or from one arm of a balanced doublet to the opposite arm in the same bit-by-bit way. As shown in the first part of this article, the distributed shunt capacitance in the cylindrical linear antenna is *not* uniformly distributed along the conductor length: It is maximum nearest the antenna input terminals, and at a minimum at the end (or ends) of the antenna.

**table 1. Mean characteristic impedance of cylindrical antennas.**

| | | |
|---|---|---|
| Doublet | $K_A = 120\,[\log_e \frac{2(h_t')}{a'} - 1]\ ohms$ | 1 |
| Grounded monopole | $K_m = 60\,[\log_e \frac{2(h_t')}{a'} - 1]\ ohms$ | 2 |

Input impedance, $Z_{in}$, of uniform characteristic impedance transmission lines (of length equal to $h_n^\circ$ where $n = 1, 2, 3 \ldots n$)

End open circuited

$$(Z_s = \infty) \qquad Z_{in} = -jK_{A,M}\ cotan\ h_n^\circ \qquad (3)$$

End short circuited

$$(Z_s = 0)\ Z \qquad Z_{in} = +jK_{A,M}\ tan\ h_n^\circ \qquad (4)$$

End terminated in complex impedance

$$(Z_s = R \pm jX)$$

$$Z_{in} = K_{A,M}\ \frac{(Z_s) \cos h_n^\circ = jK_{A,M} \sin h_n^\circ}{K_{A,M} \cos h_n^\circ + j(Z_s) \sin h_n^\circ} \qquad (5)$$

Note: when $j(Z_s) = j(R \pm jX) = jR + jjX = \begin{matrix} jR + j^2X \\ jR - j^2X \end{matrix}$, where: $\begin{matrix} +j^2 = -1 \\ -j^2 = +1 \end{matrix}$

**Useful Relationships**

| | | | |
|---|---|---|---|
| Inductance (henries) | $L = X_L/2\pi f$ | *henries* | 6 |
| Capacitance (farads) | $C = 1/2\pi f X_c$ | *farads* | 7 |
| Inductive reactance | $+jX_L = 2\pi fL$ | *ohms* | 8 |
| Capacitive reactance | $-jX_c = 1/2\pi fc$ | *ohms* | 9 |

Where $f$ is in Hz, $L$ is in henries, and $C$ is in farads

$$Q\,(3\ dB) \qquad Q = f_o/(f_{high} - f_{low}) \qquad (10)$$

When the linear antenna is too short electrically to oscillate naturally, two interesting things happen to its input impedance, $Z_{in} = R_{in} + jX_{in}$. First, that part of $R_{in}$ representing the resistive-like radiation term $R_r$ is smaller than that found in the naturally resonant antenna; that is, $R_r$ is less than the 36-ohm radiation resistance of the quarter wave, grounded monopole or less than the 72-ohm radiation resistance of the half-wave doublet.

At the same time, the input reactance, $jX_{in}$, of the electrically short antenna becomes capacitive. In the limit, when $h_t^\circ$ approaches zero degrees in length, the radiation resistance $R_r$ approaches zero ohms and $-jX_{in}$ approaches infinity. Because the electrically short linear* antenna acts like a series-resonant circuit operating on the low frequency side of resonance, we can "force" it back into electrical resonance by doing something which cancels out the reactive part of its self impedance. One way to do this is to insert a lumped inductor (loading coil) in series with the antenna conductor. The reactance, $+jX_{coil}$, of the lumped loading reactor needed to force-resonate an electrically short antenna of given length, $h_t^\circ$, will vary in magnitude, depending upon just *where* it is inserted in the antenna conductor. To investigate how the short antenna operates when the loading coil is placed *anywhere* along the antenna conductor, you must arrange the analogue transmission used to represent the antenna in such a way that it can handle any possible coil location.

**Fig. 1A** shows a *doublet* antenna of total length, $2h_t^\circ$, in the form of an analog coaxial transmission line. This is a model of the actual antenna, so think of the inner conductor of the analog line as the equivalent antenna conductor and the coaxial shield as the surrounding ground plane surface beneath the vertical monopole (or the influence of the capacitance to the other half of the doublet). The analog transmission line has a "uniform" characteristic impedance $K_A$, $K_m$, (the subscript $A$ will be used for the doublet and $m$ for the monopole). This analog transmission line $K_{A,m}$ is the mean or average value of the characteristic impedance of the cylindrical antenna which *actually varies* along its length. The mean characteristic impedance of the cylindrical antenna, given by Schelkunoff's **eq. 1** and **2** in **table 1**, forms the basis for the antenna/transmission line analog key method as it is used here.

The analog line "inner conductor" is broken in the exact center to form two balanced input terminals A-B (balanced to a fictional ground located midway between them); these are the doublet input terminals. You may view a doublet of length, $2h_t^\circ$, as being composed of *two identical monopole elements* each of length $h_t^\circ$. As was shown in the first article, the input impedance of any doublet operating in free space is just twice the input impedance of *one* of its monopole elements operated against perfect ground. Because of this, from now on you can totally ignore the part of the doublet to the left of the dashed line G-G in **fig. 1A**, knowing that once you have completed the design of a grounded monopole which meets your performance requirements, you can convert it into a doublet by merely duplicating the monopole design and placing it on the other side of G-G.

The monopole has *unbalanced* input terminals A and G ground. Moving away from the terminal A toward the right, along the inner conductor of the analog line mono-

*The term "linear" is used here to distinguish between the electric antenna and its dual, the magnetic dipole or loop antenna.

pole, a break appears in the inner conductor at a distance $h_1°$ from terminal **A**. On the other side of the gap the analog line continues on an additional length $h_2°$ to its end. The end line terminals $A_3$, $G_3$ of the analog line are open circuited; exactly the same condition prevails at the tip or end of the actual monopole antenna which the line represents. The two analog line section lengths $h_1°$ and $h_2°$ always add up to the sum $h_1° + h_2° = h_t°$. Finally, in **fig. 1A** you will notice that there are terminals $A_1$, $A_2$, $A_3$ at all locations where the analog line inner conductor is cut, and just opposite on the "shield" of the analog line are corresponding "ground" terminals $G_1$, $G_2$, $G_3$.

**Approach**: You will have to occasionally refer to the equations of **table 1** in what follows. However, we want to do something other than just solve a string of equations and become bored. To do this we will use the Smith impedance chart shown in **fig. 2**. The Smith chart is sort of a motion picture version of the famous transmission line **eq. 5** in **table 1**. One important feature of the chart is that it lets you actually "see" what goes on as an impedance "moves" along the analog line (monopole), changing its magnitude as it travels to some particular place on the line.

Comparing **figs. 1A** and **2**, note that the "output

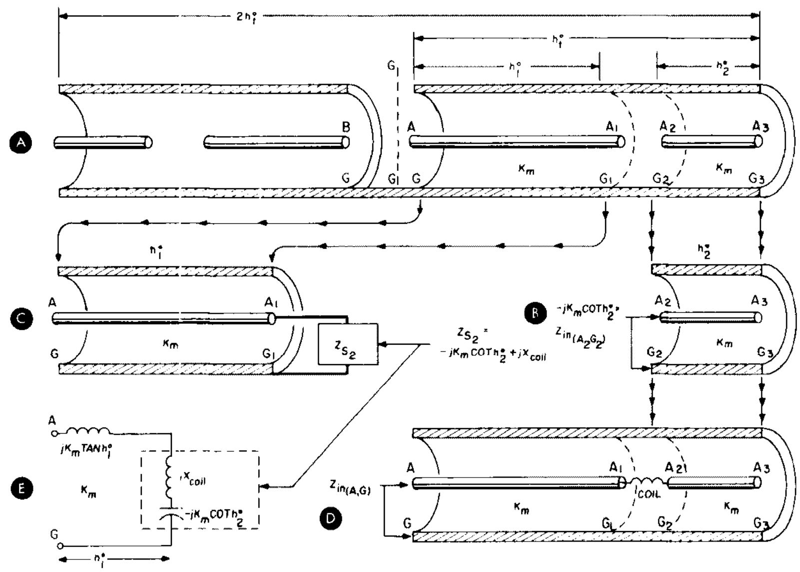



**fig. 1. (A) Analog transmission line representation of a doublet antenna of length $2h_t°$ composed of two identical monopoles of length $h_t°$. Each monopole is cut into additional length sections, $h_1°$ and $h_2°$, for insertion of the loading coil. (B) shows input impedance of line end section $h_2°$. (C) shows line section $h_1°$ terminated in load $Z_{S2}$. Short monopole with series loading coil is shown in (D). Equivalent circuit of inductively loaded monopole antenna is shown at (E).**

Actually, in advanced work, any number of gaps or line sections can be used. In the equations of **table 1**, then, where you see a length denoted $h_n°$, you may substitute $n$ equals 1, 2, 3 . . . $n$.

Why is the analog line broken in this way? It is to permit you to insert any "gadgets" such as loading coils, series capacitors, insulators, isolating parallel resonant LC "traps", and so forth which you may wish to use in your antenna designs. This is the "antenna-dissection-by-parts" technique mentioned at the end of the first part of this article. What has been said above applies to *any use* of the analog key in solving *any* antenna problem. In what follows, however, we will restrict ourselves to just two line sections, $h_1°$ and $h_2°$, to explore the electrically short, coil loaded antenna.

## design of coil loaded, electrically short antennas

**Design objective**: By applying the antenna/transmission line analog key to any electrically short cylindrical monopole antenna, you want to find the size of the loading coil needed to resonate the antenna at any *frequency, with the loading coil located at any distance* $h_1°$ from the input terminals A, G.

terminals" $A_3$,$G_3$ at the very far end of the analog line (which represent the top or tip of the monopole) are located at the bottom of the Smith chart. These output terminals are located on the very edge of the inside rim scale at a point where $R + jX$ equals infinity. Such impedance magnitude corresponds to an open circuit; that is what the antenna tip sees as a "load" impedance. The two circular scales around the outside rim of the chart are marked off in wavelengths. In what follows, you will normally use only the one labeled "Wavelengths Toward Generator" (WTG). Just below this scale is one labeled "Wavelengths Toward Load" (WTL) which permits movement in the opposite direction along the analog line. To use these scales, all you have to do is to divide the distance in degrees you wish to move by 360 degrees to obtain distance in wavelength. The "generator" referred to is the transmitter, when it is thought of as being directly connected to the line input terminals A, G.

Because the modified analog key used here deals only with reactance (the resistive $R$ part will be added later), all the impedances will be found on the very inside rim *scale edge representing pure reactance. Observe that all* inductive reactance $+jX$ is distributed around the right

hand half of the chart circle; all capacitive reactance $-jX$ is around the left hand half. There is one more important matter: At the very center of the chart is a point which represents $Z = K_m\ (1.0 + j1.0)$ *ohms*. An impedance which reaches this particular spot on the Smith chart has a resistive $R$ part equal in magnitude to that of the characteristic impedance $K_m$ of the *particular* transmission line you are dealing with in the problem, and a zero reactance part $jX$. At point P, on the other hand, the impedance is $K_m(1.0 + j1.0)$, meaning its real part $R$ and its *inductive* part $jX$ are both equal in magnitude to the characteristic impedance $K_m$ assigned to the chart.

To "enter" an actual impedance $R \pm jX$ into the Smith chart you must first divide the actual magnitudes of both $R$ and $jX$ by the $K_m$ you have *assigned* to the chart for that particular problem. After you solve your problem on the chart and wish to "remove" your answer, you merely multiply both the real and reactive parts indicated on the chart by your assigned $K_m$ value. Charts labeled in this way are said to be *normalized.* A normalized Smith chart can be used with transmission lines of *any* characteristic impedance you feel like assigning to the chart. This feature makes the normalized chart very handy to have around. Here, in working with the analog key, we will deal only with pure reactance $\pm jX$, all located on the very inside rim of the Smith chart, but in other applications an impedance point may appear anywhere on the chart.[5]

Up at the very top of the Smith chart in **fig. 2** is a point on the inside rim edge where $Z = K_m\ (0 + j0)$ *ohms*. This is "home plate" in the ball game we will play on this Smith chart "diamond." When the objective is to resonate an antenna at a desired frequency, we *always* have to reach "home plate."

## design values

Let's start right off with a specific set of conductors for the coil loaded monopole antenna. Also, for convenience, we will use a frequency, $f = 1.97 \times 10^6\ Hz = 1.97\ MHz$ in the old 160-meter ham band. Because wavelength $\lambda'$ equals $984/1.97\ MHz = 499.49\ feet$ at our selected frequency, even "short" antennas are physically large on 160 meters.

**Conductor length**: The monopole conductor length can be anything you wish as long as it is less than ninety degrees, so let's choose a total conductor length, $h_t^\circ = 33$ *degrees*. With $1.97\ MHz$ as the selected frequency, $33^\circ/360^\circ = 0.0917$ *wavelength*. At this point in the design we will deliberately *not make any distinction whatsoever between the physical and electrical length of the monopole conductor.* The reason for this "error" will be explained later on. Taking this viewpoint, the 33 degree monopole conductor length becomes: $h_t' = 499.49\ feet \times 0.0917 = 45.80\ feet$.

**Conductor diameter**: Conductor diameter may be any size you want. However, since that conductor diameter just might have some effect on the loading coil size needed to resonate the short 33 degree monopole, select three different conductor radii: $a_1 = 0.05$ *inch*; $a_2 = 0.5$ *inch*, and $a_3 = 1.5$ *inch*. To keep units the same in subsequent calculations, change these three conductor radii to feet, getting: $a_1 = 0.004$, $a_2 = 0.04$, and $a_3 = 0.125$ *feet*.

**Loading coil location, $h_1^\circ$**: Since we are exploring loading-coil placement, let's choose: base loading, center loading, and "almost" top loading. Now, by the convention given in **fig. 1**, base loading a monopole means that $h_1^\circ$ = *zero degrees*, because no break or gap would exist along the inner conductor for this case. With $h_1^\circ$ = *zero*, that makes $h_2^\circ = h_t^\circ = 33$ *degrees*. For the center loading case $h_1^\circ$ becomes $\frac{1}{2}h_t^\circ$, or in our case, $16.5$ *degrees*. This choice also makes the remaining length $h_2^\circ$ equal $16.5$ *degrees*, too, because the sum, $h_1^\circ + h_2^\circ$, *always* has to equal $h_t^\circ = 33$ *degrees*. Finally, we will explore the almost top loading case, $h_1^\circ = 32$ *degrees*, leaving $h_2^\circ$ equal to $1$ *degree*.

**Monopole mean characteristic impedance, $K_m$**. Because we are going to use three different conductor radii with the 45.80-foot conductor length, we will obtain three different values of the mean characteristic impedance, $K_m$, to represent each of the monopoles, even though $h_t^\circ = 33$ *degrees* is fixed in this case. From **eq. 2** of **table 1**, for the cylindrical monopole, first work out $K_{m(1)}$ representing the first conductor radius, $a_1 = 0.004$ *feet*:

$$K_{m(1)} = 60[(\log_e \frac{2(h_t')}{a'}) - 1] = 60[(\log_e \frac{245.8}{0.004}) - 1]$$
$$= 60[(\log_e 22900) - 1] = 60(10.04 - 1)$$
$$= 542.32\ ohms$$

Substituting the remaining two conductor radii, $a_2$ and $a_3$ into **eq. 2**, with $h_t^\circ = 45.80\ feet$,

$$K_{m(2)} = 404.18\ ohms$$
$$K_{m(3)} = 335.80\ ohms$$

**Case 1. Base-loaded monopole. Coil in series with base terminal A at monopole input ($h_1^\circ$ = *zero degrees*).** Enter the normalized Smith chart (**fig. 2**) at its "output" end terminals $A_3G_3$ located at $0.250\ \lambda$ on the WTG scale (the monopole top). Since the loading coil is located all the way down at the base of the 33-degree long monopole, we must travel that entire distance along the analog line to reach this "coil location" point. When we get there the chart will give us the normalized value of the capacitive input reactance, $-jX_{in(A,G)}$ of our monopole for this case. Since the "starting point" in the journey along the line is designated $0.250\ \lambda$ on the chart, we must *add* the distance $33^\circ/360^\circ = 0.092\ \lambda$ to that of our starting point to get the total distance:

$$0.250\lambda + 0.092\lambda = 0.342\lambda$$

Therefore, $0.342\lambda$ is the "stopping point" on the WTG scale. If you very carefully draw a straight line from the center of the chart to intersect this $0.342\lambda$ point on the WTG scale, it cuts through the $-1.54$ magnitude on the capacitive reactance scale. The rest is easy. The reactance $-1.54$ means simply $-jK_m(1.54)$ *ohms*. This is the *normalized* reactance $-jX_{in(A,G)}$ of any *cylindrical monopole* whose *electrical* length is 33 degrees. To

change this normalized capacitive reactance to represent the actual reactance $-jX_{in(A,G)}$ of the three particular monopoles under discussion, you need only multiply that normalized value by each of your three calculated $K_m$ values. But wait! For *only* the case of a loading coil located at the monopole input terminals, the size of the loading coil reactance is the *same* as the absolute magnitude of $-jX_{in}$, if you just change the sign of the reactance to a plus. You can immediately determine the reactances of all three loading coils. They are simply $+jK_m\ (1.54)\ ohms$:

$$+jX_{coil(1)} = +j\,835.10\ ohms \qquad (K_m = 542.32\ ohms)$$
$$+jX_{coil(2)} = +j\,622.22\ ohms \qquad (K_m = 404.18\ ohms)$$
$$+jX_{coil(3)} = +j\,517.13\ ohms \qquad (K_m = 335.80\ ohms)$$

It's a good thing we chose three different conductor radii; the calculations indicate that conductor diameter certainly does change the size of loading coils in an electrically short antenna of the *same* electrical length $h_t^\circ$. Not only that, but "fat" conductors need smaller loading coils than thin conductors to produce resonance in short monopoles.

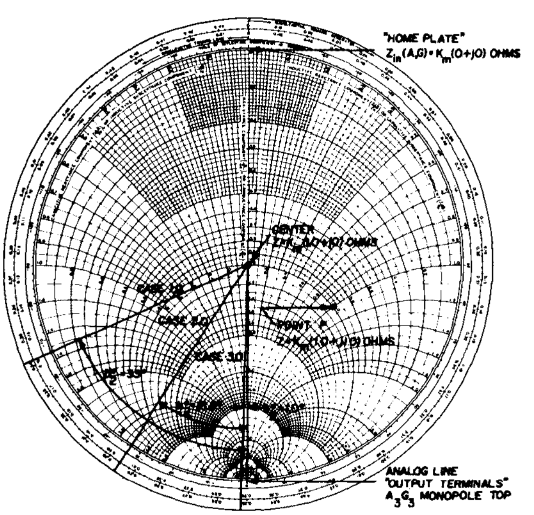



**fig. 2. Normalized Smith chart which gives solution for loading coil reactance necessary to resonate a grounded monopole ($h_t^\circ = 33^\circ$) of any $K_m$ and three locations, $h_t^\circ$, for the loading coil: Case 1, base loading; Case 2, center loading; Case 3, loading 1° from top of monopole.**

**Case 2. Center-loaded monopole**: $h_1^\circ = 16.5^\circ$ $h_2^\circ = 16.5^\circ$ Because we start at the top of the antenna/analog line, the distance to the coil location is always $h_2^\circ$. In the center loaded case, then the distance of travel will be $16.5^\circ/360^\circ = 0.046\lambda$. Adding that to the $0.25\lambda$ starting point, the stopping point on the WTG line from the center of the chart through this new point, $0.296\lambda$, you'll find a larger normalized value of capacitive reactance, $-3.38$, which you now know to be $-jK_m(3.38)\ ohms$. Before you get ahead of me and begin happily changing the sign of that reactance multiplied by the three values of $K_m$, and generating a new table of loading coils, *hold it*! The $h_1^\circ$ is no longer zero degrees! This capacitive reactance $-jK_m(3.38)\ ohms$ is really $-jX_{in(A_2G_2)}$, as shown in **fig. 1B**.

Looking over to fig. **1E**, you'll see that this "capacitor" produced by monopole analog line section of length $h_2^\circ$ is connected in series with the loading coil reactance $+jK_{coil(5)}$ and this *combination* becomes load impedance $Z_{s(2)}$. Now $Z_{s(2)}$ is seen in **fig. 1C** to be connected across the output terminals, $A_1$-$G_1$ of the lower monopole analog line section of length $h_1^\circ$. In fig. **1E**, the lower line section $h_1^\circ$ will "insert" an additional inductive reactance $+jK_m(\tan h_1^\circ)$ in series with the monopole analog line input terminals $A$-$G$. If you cancel out *all* the capacitive reactance at the loading coil gap which is produced by the upper line section $h_2^\circ$, by inserting a loading coil reactance $+jX_{coil}$ of equal magnitude, you will go "skidding past" the *home plate* at $Z_{in(A,G)} = K_m(0 + j0)\ ohms$, and end up with a $Z_{in(A,G)}$ reactive part $jX_{in}$ just equal in magnitude to this $jK_m \tan h_1^\circ$ produced by the lower monopole analog line section. What you really need is the "resonant" reactance condition

$$jX_{in(A,G)} = jK_m(X_{coil}) + (-jK_m X_{in}(A_2G_2)) + jK_m \tan h_1^\circ = jK_m(0.0), \qquad \textbf{(1)}$$

where $-jK_m X_{in(A_2G_2)}$ is really just $-jK_m \cot h_2^\circ$ from eq. 3 in table 1. **Eq. 3** is the thing we are solving with the Smith chart when we enter it at the antenna top and go traveling along the WTG scale a distance $h_2^\circ$ to the "stopping" point. When you want high accuracy, use eq. 3, because you can't read the Smith chart that closely. It is like a slide rule in this respect, but good enough for preliminary design. Looking at eq. 1 above, you can now see how to determine the size of the loading coil; just leave $+jX_{coil}$ on the left side, and put everything else on the other side of the equals sign. Then,

$$jX_{coil} = jK_m(\cot h_2^\circ) - jK_m(\tan h_1^\circ)\ ohms \qquad \textbf{(1-1)}$$

Since you just obtained $jK_m(\cot h_2^\circ) = jK_m(3.38)$ from the Smith chart, and you know $h_1^\circ$ also equals $16.5$ *degrees*, and $16.5^\circ - 0.30$

$$jX_{coil} = jK_m(3.38) - jK_m(0.30) = +jK_m(3.08)\ ohms$$

Now you can go ahead and generate a list of center loading-coil reactances:

$$jX_{coil(1)} = +j\,1670.35\ ohms \qquad (K_m = 542.32\ ohms)$$
$$jX_{coil(2)} = +j\,1244.87\ ohms \qquad (K_m = 404.18\ ohms)$$
$$jX_{coil(3)} = +j\,1034.26\ ohms \qquad (K_m = 335.80\ ohms)$$

Notice that the loading coils for the center-loaded case in the $h_t^\circ = 33\ degree$ monopole are $3.08/1.54$ or two times as large as the loading coils needed to base load the *same antenna* to the *same* frequency. Also again, "fat" conductors still require smaller loading coil size. Does this coil growth trend continue with increasing $h_1^\circ$? If

so, what is the "rate" of increase? To find out, go on to the "almost top loaded" case where $h_1^\circ$ is 32°, and $h_2^\circ$ is only 1 degree.

**Case 3. Almost top-loaded** monopole $h_1^\circ = 32^\circ$ and $h_1^\circ = 1.0^\circ$. Having to go only $1.0^\circ/360^\circ$ equals $0.002\lambda$, add that to $0.250\lambda$ and obtain $0.2520\lambda$ as the stopping point on the WTG scale. However, now you cannot read the normalized capacitive reactance $-jK_m(X_{in})A_2,G_2$ – the Smith chart reactance scale is too cramped in this region, which is approaching infinite values. No problem! Simply use **eq. 3** from **table 1** and to find $-jK_m(X_{in})A_2G_2$:

$$-jK_m(X_{in})A_2G_2 = -jK_m(\cot 1.0^\circ) = -jK_m(57.29) \text{ ohms}$$

Flipping the sign of the answer above to plus (and knowing that tangent $h_1^\circ$ = tangent $32^\circ = 0.625$) we plug these values into **eq. 1-1**:

$$+jX_{coil} = jK_m(57.29) - jK_m(0.625) = +jK_m(56.66) \text{ ohms}$$

Our final table of coil reactances follows

| | |
|---|---|
| $jX_{coil(1)} = +j\ 30730$ ohms | ($K_m = 542.32$ ohms) |
| $jX_{coil(2)} = +j\ 22902$ ohms | ($K_m = 404.18$ ohms) |
| $jX_{coil(3)} = +j\ 19028$ ohms | ($K_m = 335.80$ ohms) |

Clearly, above the center loading-coil location, coil reactance grows rapidly, at an exponential rate. At *1.97 MHz*, **eq. 6** from **table 1** gives the inductance of these coils for a location only one degree from the antenna top as

$$\text{Coil }(1) = 2.48 \times 10^{-3} \text{ henry} = 2.48 \text{ mH}$$
$$\text{Coil }(2) = 1.85 \times 10^{-3} \text{ henry} = 1.85 \text{ mH}$$
$$\text{Coil }(3) = 1.53 \times 10^{-3} \text{ henry} = 1.53 \text{ mH}$$

Coils of this size are large enough to serve as decoupling chokes in the power supply leads in your transmitter! Incidently, before turning to other matters, note that if you want to translate the short, coil-loaded monopole design technique to any *other* ham band; select an $h_t^\circ$ other than 33 degrees; or use monopole conductor radii other than 0.050, 0.50, or 1.5 inch, you will have to calculate $K_m$ again for the new case. Once this is done, you can quickly proceed as shown here to find the reactance of the needed loading coils at any height, $h_1^\circ$. Also, as you will later see, in terms of antenna efficiency it is best to reduce the size of the loading coil; the largest diameter conductor you can use will help in this respect.

Now that you have seen how the size of the loading coil changes with its placement in the monopole, you may well ask, "Why should I use a coil location other than $h_1^\circ$ equals zero degrees? Why not just base load short monopoles?" To answer these critical questions we must turn to the problem of antenna radiation resistance in electrically short antennas.

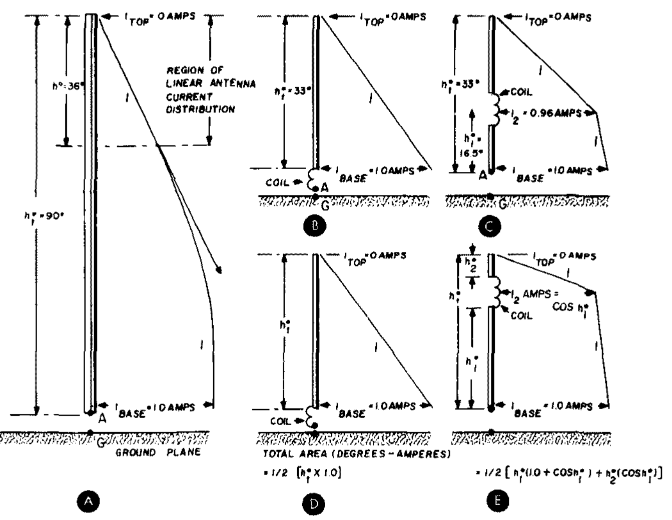



**fig. 3. (A) Sinusoidal current distribution along naturally resonant quarter-wavelength monopole antenna. (B) Linear current distribution for base-coil loaded monopole,** $h_t^\circ = 33$ *degrees*. **(C) Same height monopole with center-coil loading. (D) Total current area for base coil loaded monopole,** $h_t^\circ$ **less than 36°. (E) Total current area for short monopole antenna with loading coil at distance** $h_1^\circ$ **from base. All monopoles are assumed to be resonant.**

## radiation resistance of electrically short antennas

Up to now we have been merely plugging in $R = 0$ ohms in our antenna input impedance, $Z_{in(A,G)} = R_{in} + jX_{in}$ *ohms*. The modified antenna/transmission line analog key we've been using here gives us only the reactive part. Don't blame Dr. Schelkunoff for that. His elegant mode theory model *gives both* $R$ and $jX$ answers for $Z_{in(A,G)}$, but it employs some high power math. However, it turns out that the answers we are getting for $jX_{in}$ compare quite accurately to those obtained by his more refined equations. That's fine. But how do we get the resistive part, $R$?

To add the $R$ part to our answer, we assume $R$ equals the antenna radiation resistance $R_r$. Then we just look it up in graphs giving $R_r$ as a function of antenna length $h_t^\circ$. Such curves are found in the literature.[6] This is fortunate, because calculating radiation resistance, $R_r$,

for any antenna length $h_t^\circ$ and current distribution is not exactly child's play. So it seems we have it all wrapped up: Just look up $R_r$ for our particular antenna length $h_t^\circ$ in degrees, and plug it into $Z_{in} = R + jX$. Unfortunately, it's not quite that easy. For antennas whose length is greater than $h_t^\circ = 36$ *degrees*, the published curves give only the radiation resistance of electrically short antennas *which do not use reactive loading*. Therefore, you must be able to calculate the $R_r$ of the reactively loaded, electrically short antennas when $h_t^\circ$ is less than 36 degrees. When you do that you will see why base loading is not the optimum way to resonate electrically short antennas.

**Fig. 3A** shows the idealized, sinusoidal current distribution along a *naturally* resonant, quarter-wavelength vertical monopole antenna operating against a ground plane. If you plot the function, sine $h_t^\circ$, on graph paper, starting with zero degrees at the top of the monopole (where current in this case is zero), you will find that within the first 36 degrees you do not obtain a curve at all; for all practical purposes, you will get a straight or linear line plot. Only at lengths greater than 36 degrees does the graph plot begin to look anything like a curve. Because of this fact you can deal with the current distribution of monopole antennas, less than 36 degrees high as *straight-sided* geometric figures. Accordingly, fig. 3B shows a base-loaded monopole of height $h_t^\circ = 33$ *degrees*, with a triangular current distribution. In fig. 3C the same height antenna, center loaded, still has a straight-sided current distribution in the shape of a trapezoid.

Now, if you measure antenna section heights $h_1^\circ$ and $h_2^\circ$ in electrical degrees, and the current amplitude in amperes, the total area of the current distribution can be expressed in units of degree-amperes. One of the cruder (but good enough ways) to explain radiation resistance is to say that Nature only "sees" the antenna's total "exposed" current area. Based on this engineering "viewpoint" you can write an expression* which gives the radiation resistance, $R_r$, of electrically short ($h_t^\circ$ less than $36^\circ$) monopole antennas operating over a ground plane as

$$R_r = 0.01215\,A^2 \qquad \textbf{(1-2)}$$

where $A$ is the total exposed antenna current area in degree-amperes.

For a doublet of length $2h_t^\circ$ in free space, composed of two short, reactively loaded monopoles of length $h_t^\circ$, just double the $R_r$ magnitude calculated from **eq. 1-2**. **Fig. 3D** provides a formula for finding the triangular shaped, degree-ampere area of base-loaded monopoles over ground; **fig. 3E** gives the formula for the case where the loading coil is located a distance $h_1^\circ$ from the input terminals $A$-$G$. It is important to note that, in all antennas shown in fig. 3, the antenna base current is *always* one ampere. This is the relative current amplitude which you must use because the radiation resistance of an antenna does not change with the different amplitudes of actual currents when you vary transmitter input power. For reasons which space does not permit me to go into here, when the antenna is *resonated* by a loading coil placed at a distance $h_1^\circ$ from the base input terminals, the *relative* current $I_2$ at the coil position has an amplitude of $I_2 = cosine\ h_1^\circ$.

*Eq. 1-2 is based on the theory of monopole moment.

Using the above information, you can now see what happens to radiation resistance in short antennas as $h_1^\circ$ is changed while antenna height, $h_t^\circ$, is fixed. For example, consider your antenna, $h_t^\circ = 33$ *degrees* when it is base loaded. From fig. 3D, the current area is then

$$A = \tfrac{1}{2}\,(33^\circ \times 1.0\ ampere)$$
$$= 16.5\ degree\text{-}amperes$$

From **eq. 1-2**:

$$R_r = 0.01215\,(16.5\ degree\text{-}amperes)^2 = 3.3\ ohms$$

For the case where the loading coil was moved to $h_1^\circ = 16.5^\circ$, to center load it, $I_2 cosine\ 16.5^\circ = 0.96$ *amperes relative*, giving us the area from fig. 3E as,

$$A = \tfrac{1}{2}[16.5^\circ\,(1.0 + 0.96) + 16.5^\circ\,(0.96)]$$
$$= \tfrac{1}{2}[32.34 + 15.84]$$
$$= \tfrac{1}{2}[48.18] = 24.10\ degree\text{-}amperes$$
$$R_r = 7.06\ ohms$$

You can see that center loading *increased* the radiation resistance by a ratio of *7.06/3.3 or 2.14 times*. The ohmic loss produced by a radial wire ground plane, insulator leakage, soil current resistance, and the like, might have a realistic ohmic resistance magnitude of 10 ohms. Antenna radiation efficiency, $N$, is related to the ratio between antenna radiation resistance, $R_r$ and total environmental ohmic loss, $R_\Omega$, as

$$N = \left(\frac{R_r}{R_r + R_\Omega}\right) 100\ (per\ cent) \qquad \textbf{(1-3)}$$

With ohmic loss equal to 10 ohms, the base-coil loaded monopole ($h_t^\circ = 33^\circ$) would yield an efficiency of,

$$N = \left(\frac{3.3}{3.3 + 10.0}\right)100 = 24.8\ per\ cent$$

The same height monopole, when center loaded, would yield

$$N = \left(\frac{7.06}{7.06 + 10.0}\right)100 = 41.4\ per\ cent$$

Not only is more input power radiated from the center-loaded monopole, but since $Z_{in} = R_r + R_\Omega + jX$, the magnitude of input impedance at resonance is increased, which makes impedance matching with a base network less of a headache. Right about here someone is going to remember the case where the loading coil was only 1 degree down from the top. Using the relations given, $R_r = 10.93$ *ohms*. This looks promising — increased $R_r$ means increased efficiency, right? I am sorry I must dash cold water over this happy discovery. If you could raise the current-degree ampere area of the 33-degree monopole to the indicated total area *without introducing additional ohmic loss*, you could increase

efficiency because $R_r$ would increase toward its theoretical limit. The theoretical limit in reactively-loaded short monopole (or dipole) antennas is *four* times that of the base-loaded $R_r$ (or doublet feedpoint loading), and is based on a perfectly rectangular current distribution. However, you can't approach this limit except in very short antennas, and *never* in the case of any *coil-only* loaded short antenna.

The words "without introducing any additional loss", however, lets the air out of this little balloon. Remember the rapid increase in loading-coil reactance (and thus inductance) as it climbed to higher and higher $h_1°$? You might think a loading coil is just hacked out of any commercial length of coil stock you have laying around the shack. Not true. An antenna loading coil must be tailored to your radiating system if you want to obtain maximum efficiency.

## loading coils in electrically short antennas

Space limitation prevents me from going too deeply into the critical matter of loading-coil design. Still, being terse, I'll try to cram in some vital facts. By now your intuition must tell you that loading coil loss should be kept to the barest possible minimum. That clearly means use of a high value of coil Q. Unfortunately, coil ohmic loss $R_L$ is related to coil Q by the expression

$$R_L = \frac{X_{coil}}{Q} \text{ ohms} \qquad \textbf{(1-4)}$$

Eq. 1-4 says that, as coil reactance grows with increasing $h_1°$, coil Q must be increased to retain this minimum $R_L$ coil loss. Here, also, is a sobering thought: the coil shape factor for maximum Q in a single layer coil is represented by a coil *diameter twice that of coil length.*[7] This ratio is not especially critical, but you should stay pretty close to it. So you see that as reactance increases, demanding more coil turns in increasing length, the physical diameter of the loading coil keeps increasing as it climbs the short antenna. Another important matter is the insulation holding the turns to the correct pitch. The material used for coil insulation should have the lowest possible dielectric loss factor, and a *minimum* amount of it should be employed. Moisture from the weather should be kept from the exposed loading coil by a sealed dielectric housing, because the slightest moisture film-bridging between turns, across the insulation, will reduce coil Q. Because of their low value of radiation resistance, the input current to short antennas is considerably larger than it is in, say, a naturally resonant quarter-wavelength monopole at the same input power. The antenna input (base) current, in terms of its actual magnitude, is

$$I_{in} = \sqrt{\frac{Pi}{R_r + R_o}} \text{ amperes} \qquad \textbf{(1-5)}$$

where $Pi$ is power input in watts.

Consequently, provisions must be made to handle the calculated current amplitude in the coil conductor.

*Multiply actual $I_{in}$ by cosine $h_1°$ to find $I_c$ when the coil is located at a point higher than the antenna's base-input terminals.

Voltage stress (voltage drop) across the loading coil is given by

$$V_{coil} = (I_c \times X_{coil}) \text{ volts} \qquad \textbf{(1-6)}$$

Where $I_c$ is the current flowing through the coil at its location $h_1°$ in the antenna.* If $I_c$ is in rms amperes, multiply $V_{coil}$ by *1.414* to get peak volts of stress.

Practical antenna engineers, adding this all up, feel that the center-loaded case is about optimum in terms of a trade off between increase in radiation resistance to offset environmental ohmic loss, and design of the loading coil within practical and economic limits.

A final word about tuning a short, coil-loaded antenna: Once you have achieved the high Q coil you need to resonate the little monster to the frequency in the band you select, don't try to change its inductance by means of tapped turns, turn sliders, or shorted-out turns to tune the antenna around the band. The resulting change in coil shape factor, and the eddy currents which *will circulate in the unused* turns, will *murder coil Q.* Instead, in fixed site home station versions retune the impedance matching base network you have to use anyway to lower transmission line vswr. In electrically short, mobile antennas use a telescoping section of the antenna conductor ***above*** the loading coil to establish resonance of the antenna at frequencies above or below $f_o$. Finally, I hope that you will now look at the long, skinny loading coils wound on massive insulating forms which you may see used in some short antenna designs with a healthy bit of skepticism. However, with use of the analog key, you can now design any coil-loaded, short antenna which comes to mind.

## calculated coil size accuracy

Real antennas of any kind are *not* lumped circuits. Their extensive fields reach far out into the so called near-zone region surrounding them to "feel and sense" their electromagnetic environment and to react to it. They do this by automatically adjusting their electrical characteristics to always satisfy Nature's laws of *minimum energy balance*. Electrically short antennas are the most sensitive of all, due to the much larger storage of energy within their near-zone region. To an antenna, each station site, each mobile vehicle, differs considerably from another.

It is due to this peculiar nature of the antenna, in contrast to that of lumped circuits, which makes it impossible for man to devise a calculation method which can *exactly* predict in the design stage all the changes which will occur in a real antenna operating in some actual electromagnetic environment. Therefore, working antenna engineers make the best calculations they can at the office desk, then put on their hard hats and go out to the antenna measurement laboratory and prune their creations exactly to size. This problem is least at microwave frequencies, and at its worst at frequencies in the low- and high-frequency bands. In the case of the electrically short antenna, this means pruning calculated coil size. But you can use an "insurance policy" here because first, you don't want to start out with a coil which is too small, and secondly, you don't want to do too much

pruning. How does the analog key satisfy these two design requirements for the practical designer?

Recall that when you started to select antenna conductor length for the electrically short antenna you did not make a correction between conductor physical length and its electrical length. I said we would regard them at the start as equal to one another. It turns out that unless an antenna conductor is of zero diameter, its electrical length is always greater than its physical length. If you examine eq. 3 in table 4, you will realize that if such is the case, all the magnitudes of capacitive reactance calculated from it, or its solutions via the Smith chart, are somewhat smaller than those obtained. That means that all of the calculated coil reactance values are too large.

In matter-of-fact, their size error on the plus side of tolerance is in inverse proportion to the mean antenna characteristic impedance $K_m$ you calculated them to match. Thus, the coil calculated for $K_m = 335.8$ *ohms* would resonate the 45.8 foot tall monopole to a frequency below *1.97 MHz* in the 160 meter band; the coil for $K_m$ = 542.3 ohms would resonate the same height monopole closest to, but still a bit below *1.97 MHz*. Not too much below *1.97 MHz* (or whatever frequency you designed for) but enough that you may safely prune the coil *up in frequency* to resonate in your own particular environment. I do that with a grid-dip oscillator which is *loosely* coupled to the grounded coil/monopole combination.

At the end of our next example, you will see how to make the correction between conductor's physical and electrical length when you need to. This, however, will be in working with naturally resonant antennas which are least sensitive to environment.

## frequency bandwidth of antennas

Design objective. You want to design a naturally resonant antenna in such a way that the vswr on the transmission line feeding the antenna does not exceed a specified maximum value at any frequency within the limits $f_{high} - f_{low}$ of a given amateur band.

Such antenna design will provide you with an antenna into which a modern transmitter, using a pi-network tank circuit, will easily load full power at any spot in the band. Most amateur transmitters are designed to load full power only into a transmission line vswr of 2.8:1 or less.

Foreward to problem. Remember that I said that the input reactance, $jX_{in}$, of an antenna of fixed length changed far more rapidly in magnitude with a change in transmitter frequency than the resistive part, $R_{in}$. The resistive part of antenna input impedance, in fact, changes so little over the frequency width $f_{high} - f_{low}$ of any presently assigned high-frequency amateur band that we may make an engineering *approximation* and view $R_{in}$ as a constant over the frequency width $f_{high} - f_{low}$.

With $R_{in}$ assumed constant over the frequency width of any amateur band, let's take up the case of a naturally resonant quarter-wavelength monopole antenna, remembering that we can always combine two such monopole designs into a single half-wave doublet if we wish.

The radiation resistance, $R_r$, of a naturally resonant quarter-wavelength monopole over perfect ground is equal to 36 ohms. If we select the natural resonant frequency, $f_o$, of the monopole to be that of the band center, then its input impedance at, $f_o$, will be $Z_{in(A,G)} = 36 + j0$ *ohms*. When we tune the transmitter vfo above and below $f_o$, the linear monopole antenna will exhibit the impedance characteristics of a series-resonant circuit. That is, it will exhibit a $-jX_{in}$ capacitive reactance in series with $R_r$ at frequencies below that of $f_o$, and a $jX_{in}$ inductive reactance at frequencies above $f_o$.

When the absolute magnitude of the reactive part of the antenna input impedance becomes just equal to the magnitude of the resistive part, antenna designers refer

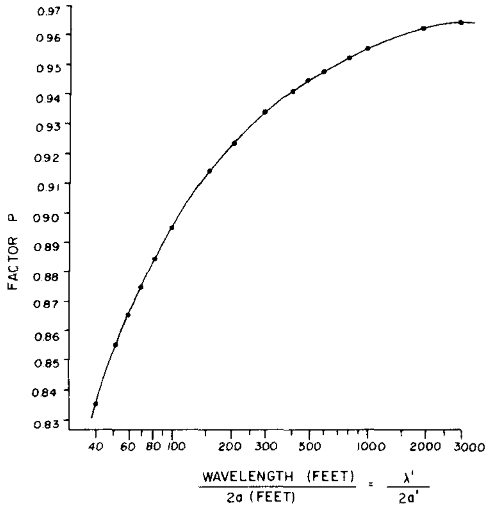



**fig. 4. Correction to obtain height, $h_t$ (feet) equal to 90 electrical degrees at resonant frequency. To use this chart use the following procedure:**

1. **Calculate wavelength from $\lambda = 984/f\,(MHz)$.**
2. **Divide monopole conductor diameter, $2a$ (feet) into $\lambda$ (feet).**
3. **Enter graph to obtain length correction factor, $P$.**
4. **Use $246P/f_o$ (MHz) to obtain resonant monopole length, $h_t$ (feet).**

**For example, what is the correct height at 3.6 MHz for a monopole constructed of 3-inch (0.25 foot) aluminum pipe?**

$\lambda = 984/3.6 = 273.33$ *feet*

$\lambda/2a = 273.33/0.25 = 1093.33$

**From the graph, $P = 0.956$. Therefore, correct monopole length is** $(246 \times 0.956)/3.6 = 65.3$ *feet.*

to the total frequency span between the input impedance values as the 3-dB or half-power bandwidth of the antenna. This impedance behavior looks like this:

$$f_{low} = 36 - j36 = 50.91 \angle -45^\circ$$
$$f_o = 36 + j0 = 36.00 \angle 0^\circ$$
$$f_{high} = 36 + j36 = 50.91 \angle +45^\circ$$

Following good practice, let's select a transmission line for our monopole antenna whose uniform characteristic impedance, $Z_o$, is equal to that of the radiation resistance, $R_r$. This calls for a coaxial line of $Z_o = 36$ *ohms* (two 72-ohm coax lines connected in parallel). When we do this, at the resonant frequency at the band

center, the vswr on the feedline will have a value of 1:1 because it is matched to its load impedance. When we tune the vfo to some frequency limit above or below resonance, $Z_{in} = 50.91 \angle \pm 45^{\circ}$ *ohms*.

At the two frequencies on either side of resonance where this impedance value is reached, the vswr in the feedline will have climbed to 2.6:1. Since the point where the absolute value of the reactive and resistive parts are equal, $|jX| = |R|$, marks the limits of the 3-dB bandwidth of the antenna, it is easy to wrongly assume that a vswr of 2.6:1 means that 3 dB or half the input power is reflected due to mismatch. However, a 2.6:1 vswr represents a mismatch condition where only 0.97 dB of the incident power is reflected. The reason for this much lower reflection is based on an important electrical law, Thevenin's theorem.*

The actual frequency width, $f_{high} - f_{low}$, at which the vswr rises to 2.6:1, however, depends upon the electrical nature of the antenna. To find out how to solve this problem, first write the quality factor Q for an antenna in terms of the 3 dB bandwidth as

$$Q = \frac{f_o}{f_{high} - f_{low}} \qquad \textbf{(1-7)}$$

All you need to solve the problem is a single design parameter of the antenna which is related to Q in frequency width. As you probably have guessed by now, this needed antenna design parameter is the mean characteristic impedance $K_m$ of the monopole antenna. It is related to frequency Q as,

$$K_m = (46.5\ Q) + 3 \ ohms \qquad \textbf{(1-8)}$$

Once you know these relationships, you can take the following design steps: First plug your particular amateur frequency limits, $f_{low} - f_{high}$, and $f_o$, into **eq. 1-7** to find Q; then plug the obtained Q value for the band into **eq. 1-8** and determine the needed $K_m$ of the monopole. Then, all you have to do is solve for the antenna conductor length-to-radius ratio to give you this value of $K_m$.

To obtain this conductor size ratio, rewrite **eq.** 2 of **table 1** in the following way,

$$\frac{K_m}{60} + 1 = \log_e \frac{2(h_t')}{a'} \qquad \textbf{(1-9)}$$

Let's try this method out, first for the 10-meter band, and then for 80 meters. For 10 meters, the bandwidth limits are *29.7 MHz* to *28.0 MHz*. The band center of 28.85 MHz is our $f_o$. Solving for Q from **eq. 1-7**

$$Q = \frac{28.85}{29.7 - 28.0} = 16.97$$

$$K_m = (46.5 \times 16.97) + 3 = 792 \ ohms \qquad \text{(from eq. 1-8)}$$

$$\frac{792}{60} + 1 = \log_e \frac{2(h_t')}{a'} = 14.202 \qquad \text{(from eq. 1-9)}$$

The needed conductor length-to-radius ratio is simply the natural antilogarithm obtained from **eq. 1-9**, or

$$\frac{2(h_t')}{a'} = 1.472 \times 10^6$$

$$a' = \frac{2(h_t')}{1.472 \times 10^6}$$

Here you need $h_t'$ for the band $f_o$. Again start by *first assuming* that $h_t' = 0.250\lambda$ at 28.85 MHz (no correction as yet). Then $h_t' = 0.250 \times 34.11' = 8.53$ *feet*; $2(h_t') = 17.05$ *feet.*
Therefore,

$$a' = \frac{17.05'}{1.472 \times 10^6} = 1.16 \times 10^{-5} \ feet$$

$$= 1.39 \times 10^{-4} \ inches$$

*conductor diameter* $= 2.78 \times 10^{-4}$ *inches*

This is like shooting down a butterfly with a cannon! Here you went to some effort only to find that any gauge copper wire, physically strong enough to be suspended as an antenna, will give you the needed frequency vswr bandwidth on 10 meters. All this fancy frequency bandwidth design isn't for amateurs, right? To check on that, knowing how touchy Mother Nature can be, let's try the same exercise on the 80-meter band. It's width is 4.0 MHz to 3.5 MHz, with 3.75 MHz in the center. So

$$Q = \frac{3.75}{4.0 - 3.5} = 7.5$$

$$K_m = (46.5 \cdot 7.5) + 3 = 351.75 \ ohms$$

$$\frac{351.75}{60} + 1 = \log_e \frac{2(h_t')}{a'} = 6.86$$

Antilog (*e*) of 6.86 = 955.75

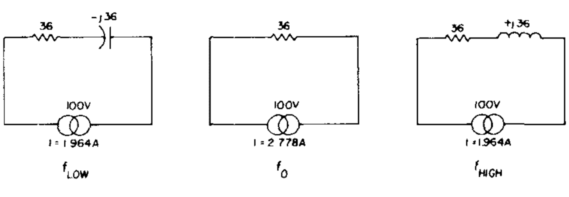
$$h_t' \ (0.250\lambda) \text{ at } 3.75 \text{ MHz} = 65.6 \ feet$$

$$a' = \frac{131.2'}{955.75} = 0.137 \ feet = 1.65 \ inches$$

*conductor diameter is 3.29 inches*

This is a horse of another color! You need an antenna

*Basically, Thevenin's theorem states that the current in any impedance connected to two terminals of a network is the same as if that impedance were connected to a generator whose voltage is equal to the open-circuited voltage at the terminals of the network. This can be illustrated by feeding an antenna with a constant-voltage source as shown below:




With a 100-volt constant-voltage source in series with the impedance, the current flow is 1.964 amp at $f_{low}$ and $f_{high}$, and 2.77 amp at $f_o$. *The power dissipation in the real part of the impedance at the three frequencies is as follows:*

| | |
|---|---|
| Power at $f_{low}$ | = *138.86 watts* |
| Power at $f_o$ | = *277.72 watts* |
| Power at $f_{high}$ | = *138.66 watts* |

Note that the power difference is exactly 3 dB.

conductor whose minimum diameter is 3.3 inches to use in an antenna which will cover the entire 80-meter band without exceeding 2.6:1 vswr in the feedline. This means a length of 3.5 inch aluminum irrigation tubing; or triangular, but uniform diameter, triangular cross section tower at least 3.5 inch on each side for the 80 meter monopole.

For a doublet we can simulate the needed conductor diameter in the form of a lightweight cage design using a minimum of eight wires of say no. 18 AWG Copperweld. These wires can be arranged equally around the perimeter of circular plastic formers spaced along the length of each half of the doublet. At both ends of the dipole sections all wires should be brought to a point to form a taper where the insulators are attached. This is for reduction of shunt capacitance across the doublet input terminals. Excess shunt capacitance, either from the monopole conductor base to ground — or across the balanced doublet input terminals — will produce a non-symmetrical vswr curve at frequencies equally spaced above and below resonance. Reduction of base shunt capacitance is the reason you see those long tapers at the base of large cross section, vertical tower monopoles used by broadcast stations.

Finally, I promised to give a correction factor to take in the physical length of the monopole conductor, as a function of its diameter, so that it is ninety electrical degrees long at the resonant frequency. **Fig. 4** gives the correction curve and the detailed procedure needed to do this. When you actually design and put one of these naturally resonant, broad-banded antennas on the air, you will find that the vswr at the band edges will be somewhat less than 2.6:1. This comes about because of the addition of environmental ohmic losses induced in the antenna input impedance by the antenna site, as well as from the small decrease in antenna $K_m$ when you shorten the conductor to resonance by use of *the data in* **fig. 4.**

In this article I have only scratched the surface in applying the analog key to the field of practical antenna design; to do a thorough job would require a book on the subject. Enough material has been given here, however, to enable you to quickly become proficient in the use of the analog key and to turn it loose on your own pet ideas for antenna design.

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**ham radio**

# simple broadband antenna for 10 GHz

## The coax-to-waveguide adapter provides an excellent broadband antenna for amateur microwave work

**Interest in the 10-GHz amateur** band is growing rapidly as the lower vhf and uhf bands are filling with fm repeaters. However, getting started on 10 GHz is difficult due to the absence of commercially available rigs;* this forces amateurs to resort back to home-brewing their own equipment. One problem is antennas. Having built several antennas for the 3 cm band, including polyrods, helix, dish, horn and even a Yagi beam, we have found that, generally speaking, the simpler, the better. The simplest and most reliable antenna we used is a Hewlett-Packard model X281A coax-to-waveguide adapter. Similar results can be had with adapters from other manufacturers. However, the Hewlett-Packard unit was available and covered the 3 cm amateur band.

Fig. 1 shows the vertical radiation pattern of the H-P X281A, and fig. 2, the horizontal radiation pattern. Waveguide antennas are like slot antennas in that the electric vector or polarization vector is perpendicular to the wide dimension of the waveguide. The vertical half-angle beamwidth is 65° and 80°, respectively, at the 3-dB and 10-dB points. The horizontal half-angle beamwidth for the 3-dB and 10-dB points is 35° and 45°. The horizontal beamwidth is narrower than the vertical

*An exception is the recently announced 10-GHz *Gunnplexer* available from Microwave Associates (see page 86, March, 1977, and page 10, April, 1977, *ham radio*).

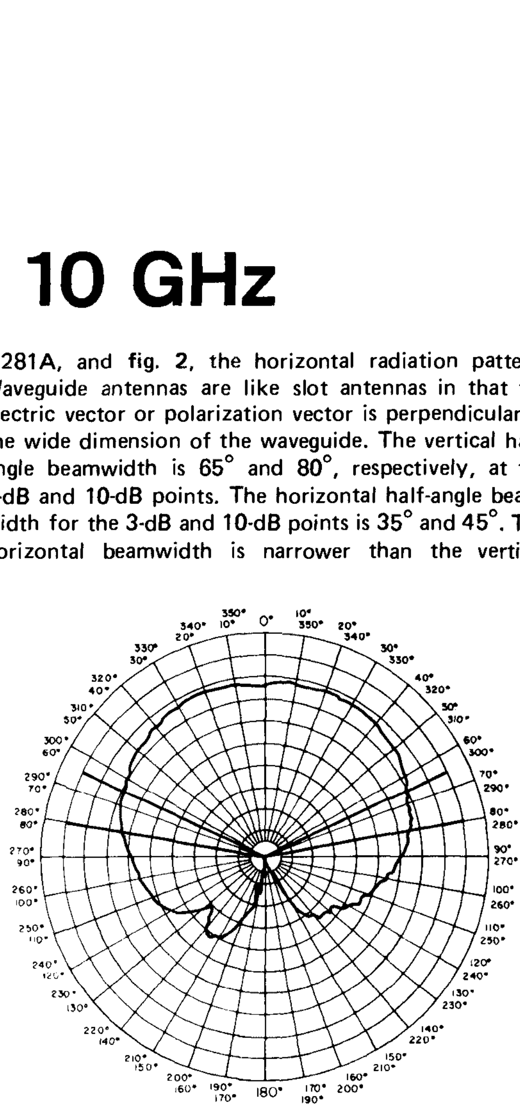


**fig. 1. Vertical radiation pattern for the Hewlett-Packard X281 coax-to-waveguide adapter measured at 10.25 GHz. The 3-dB half-angle beamwidth is 65 degrees; 10-dB half-angle beamwidth is 80 degrees.**


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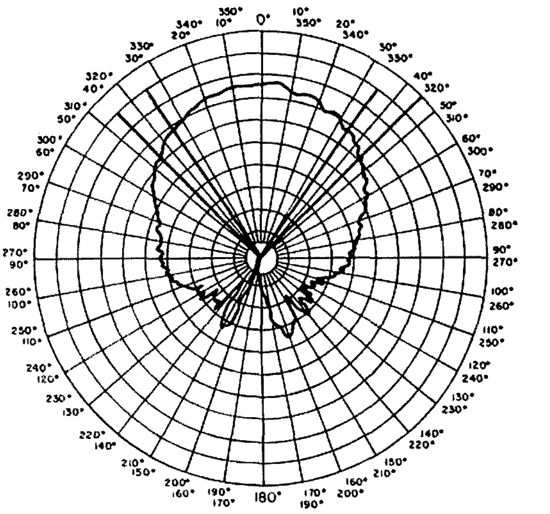


**fig. 2. Horizontal radiation pattern of the Hewlett-Packard X281 coax-to-waveguide adapter measured at 10.25 MHz. The 3-dB half-angle beamwidth is 35 degrees; 10-dB half-angle beamwidth is 45 degrees.**

because the horizontal antenna dimension is larger than the vertical dimension.

The reason for measuring the 3-dB and 10-dB points is that, when the antenna is used as a feed for a parabolic reflector, we use the 10-dB beamwidth; the 3-dB beamwidth is quoted when the antenna is used by itself. Plotted in fig. 3 is a graph of the optimum feed 10-dB half-angle beamwidth as a function of the ratio of the reflector's focal length to diameter ratio or *f* number. It is apparent that the coax-to-waveguide adapter works best when feeding parabolic reflectors which have *f* numbers between 0.68 and 1.25.

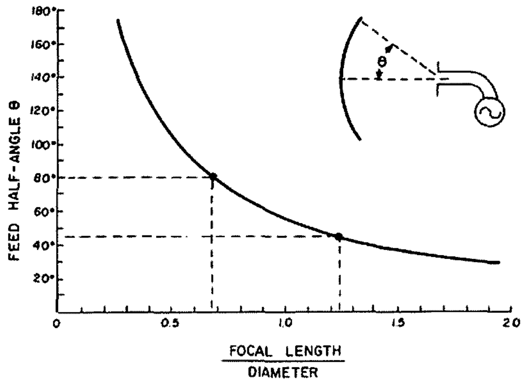


**fig. 3. Optimum feed half-angle beamwidth vs focal length to diameter ratio (*f* number) of a parabolic reflector. The simple antenna described here should be used with reflectors with *f* numbers between 0.68 and 1.25 as shown by the dashed lines.**

In conclusion, whether used directly as an antenna or used in conjunction with a parabolic dish, the simple, broadband coax-to-waveguide adapter provides a low-cost, dependable microwave antenna.

**ham radio**

# automatic position control for the HAM-M rotator

## IC logic provides automatic brake release and positive position control for this popular antenna rotor

When I built my cubical quad antenna, I selected the Ham-M rotor because of its rugged construction. A particularly important feature that influenced this decision was the brake, which protects the gear drive from damage in a wind storm. The brake, however, has the disadvantage of engaging before the antenna has stopped turning, which puts a severe torque on the tower. I felt it was easier to design a brake delay circuit than to find a way to reinforce my tilt-over mast.

Having decided to modify the brake circuit, it was natural to see if additional improvements could be made to the Ham-M control circuit. The result of my analysis was that it would also be beneficial to have an automatic position control, so I designed the following circuit to provide these features.

### principle of operation

The control concept is shown in fig. 1. A regulated power supply drives a bridge circuit consisting of R8 and R9. R8 is the existing position-sensing pot in the rotator with R9 being in the control box. When the rotor is in the desired position, voltages on each of the pot wipers will be equal. If you turn R9, the voltages will no longer be equal. The voltage difference is proportional to the position error. This error voltage is amplified by error amplifier U2. U2 output is checked by comparators U3 and U4, which determine if the rotor is to turn to the left, turn to the right, or stay in place. Logic signals $\overline{\text{LEFT}}$ and $\overline{\text{RIGHT}}$ go to a timing circuit, which drives the relays. The relays in turn, actuate the rotor motor and brake release coils.

As the rotor turns, R8 is turned. When the desired position as set on R9 is reached, the error voltage goes to zero. The motor is then turned off, and after a short delay, the brake is engaged. This action occurs automatically.

### circuit description

Fig. 2 is the schematic of the error amplifier and comparators. Since the wiper of R8, the position-sensing pot, is grounded to the rotor housing, a floating power supply is needed. This supply is provided by meter transformer T1 and voltage regulator U1. The meter circuit is the same as in Ham-M design.

**Error amplifier.** Error amplifier U2 is a common 741 op amp. Any voltage difference between pins 2 and 3 is amplified by 10,000 times or more. This gain is reduced by negative feedback through R11 and R12. The overall effect is that U2 output is 1 volt for every 6 degrees of error in position.

**Comparators.** U3 and U4 are LM311 comparators. These devices are similar in function to the op amp except that the output is a digital signal. If pin 3 is more positive than pin 2 the output is grounded; otherwise, the output is pulled up to the +12-volt supply through the external components.

U3 pin 2 is connected to a threshold voltage, which can be adjusted between 0.5 and 2 volts. If U2 output is greater than this threshold voltage, U3 output will be

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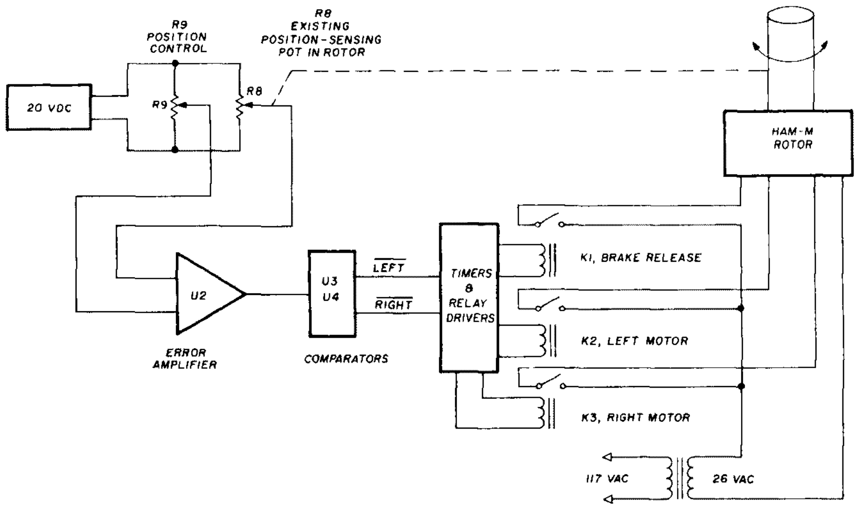


**fig. 1. Simplified block diagram of the automatic position control system for HAM-M antenna rotor.**

low, indicating that the rotor should turn to the right. As the rotor turns, U2 output decreases until U3 output again goes high. U4 works in a fashion similar to U3 when the antenna is turning to the left. R26 and R27 provide positive feedback, and hence a little hysteresis, to the comparator action so that no uncertainty will occur when the inputs are nearly equal.

There's a small range in the error amplifier output where neither comparator output is low. This results in a dead band in the control action, which is unavoidable for two reasons. First, there's no simple means of gradually reducing motor speed so that the rotor stops exactly in the right place. (SCR circuits are well known rfi generators.) Furthermore, the brake locks in incre-

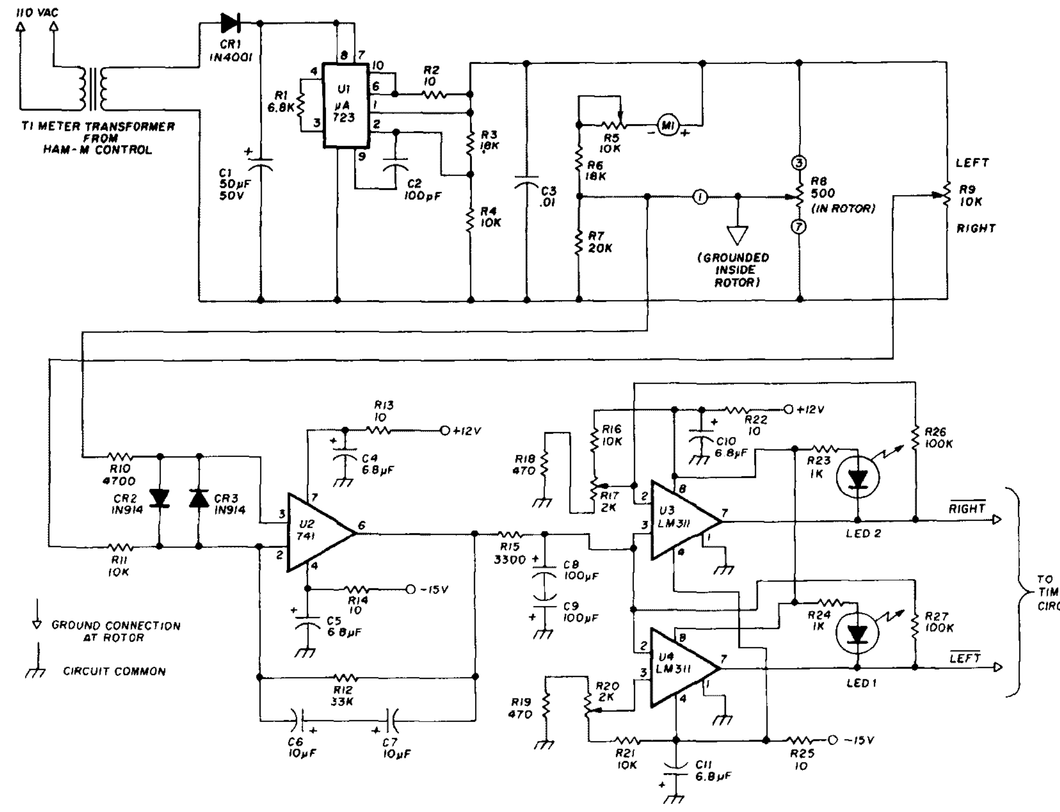


**fig. 2. Error amplifier and comparator schematic.**

ments of 3° 45′, which is of little practical concern since few antennas have extremely narrow beamwidths.

The bypass capacitors and decoupling resistors in the power-supply leads of U2, U3, and U4 are a precaution against instabilities that sometimes bother IC circuits. This part of the circuit design is possibly overcautious, but I felt it was better to be safe.

Because the wires from the position-sensing pot, R8, are in the same cable as the wires supplying power to the motor, a considerable 60-Hz signal is superimposed on the position signal. C6, C7, C8, C9, and R15 filter this hum before it reaches the comparators. The hum is further reduced by not grounding the circuit common inside the control box. However, safety and rfi considerations require that the cabinet be properly grounded.

**Timing circuit.** The timing circuit, fig. 3, is based on the popular NE555 timer IC. A low input on pin 2, from either the comparator or the manual test switch, will start the timing sequence. U5 discharges C13 through pin 7. Normally C13 could begin to recharge through R29; but in this circuit, Q2 keeps C13 discharged as long as the input is low. Meanwhile the output, pin 3, is high, which causes CR3 to conduct, energizing brake-release relay K1. Simultaneously Q3 is also turned on. Q3 then turns on K2, which causes the rotor to turn to the left. U5 output also acts through Q4 to keep U6 reset. U6 cannot start timing until U5 has timed out.

When U5 input goes high, Q2 and Q3 turn off; hence the motor is turned off. C13 then begins to charge through R29. When C13 reaches 8 volts, U5 turns off and the brake is engaged.

If, during the coasting period, you wish the antenna to rotate further, Q2 and Q3 can be turned on immediately and the motor will be reenergized. If you wish to reverse rotation direction, it will be necessary for U5 to time out before U6 can be set and K3 energized to turn the rotor to the right. In short, an operator error cannot cause the circuit to jam.

**Power supplies.** This control circuit requires three dc power supplies. These are a 20-volt, 60 mA supply for the positioning bridge and meter; (**fig.** 2) a +12-volt 300 mA supply for the ICs and relays; and a −15 volt supply for the error amplifier and comparators. The +12-volt and −15-volt supplies are shown schematically in **fig. 4.**

The design is based on the $\mu$A723 regulator. It would probably be easier and less expensive to use the newer

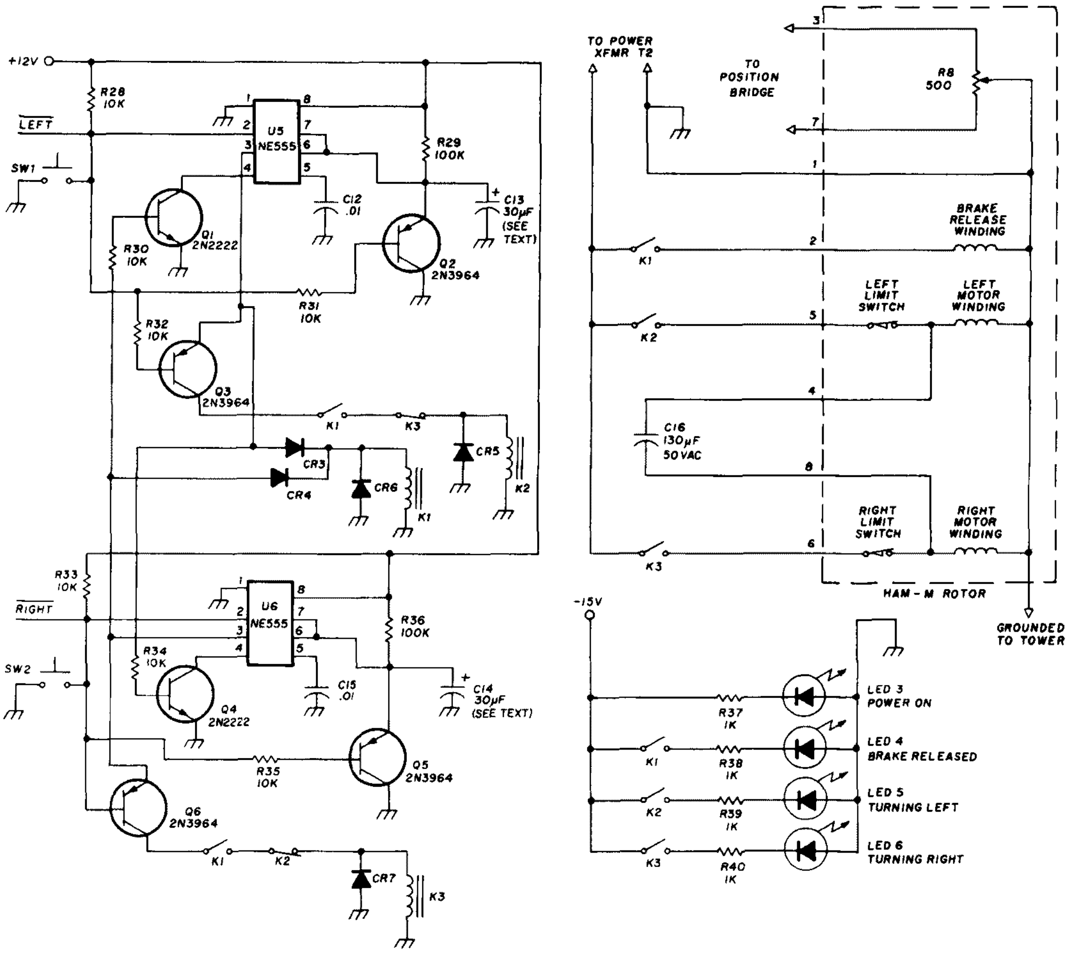


**fig. 3. Timing-circuit schematic. C13, C14 see text. CR3-CR7 1N645A or similar. K1-K3 12 Vdc 185-ohm coil, 29 V 5A contacts (Potter & Brumfield R10-E1-X4-V185 or Allied TF154-CCCC-12V). LEDs are HP 4403.**

three-terminal regulators such as the LM320 and LM340 series. Because there are few, if any, 20-volt regulators available, the position-bridge circuit could be operated at 18 volts. Whichever voltage you choose, the supply should be fused at no more than 1/8 ampere unless the regulator has internal current limiting.

Nothing is critical about the control-circuit assembly. I removed the transformers, meter, and motor capacitor from the Ham-M control box and put everything into a larger cabinet. Alternatively, the transformers and meter can remain in the Ham-M box and the new circuit put into a separate enclosure. Either way, most of the wiring can be done with a Vector wiring pencil or on a PC board.

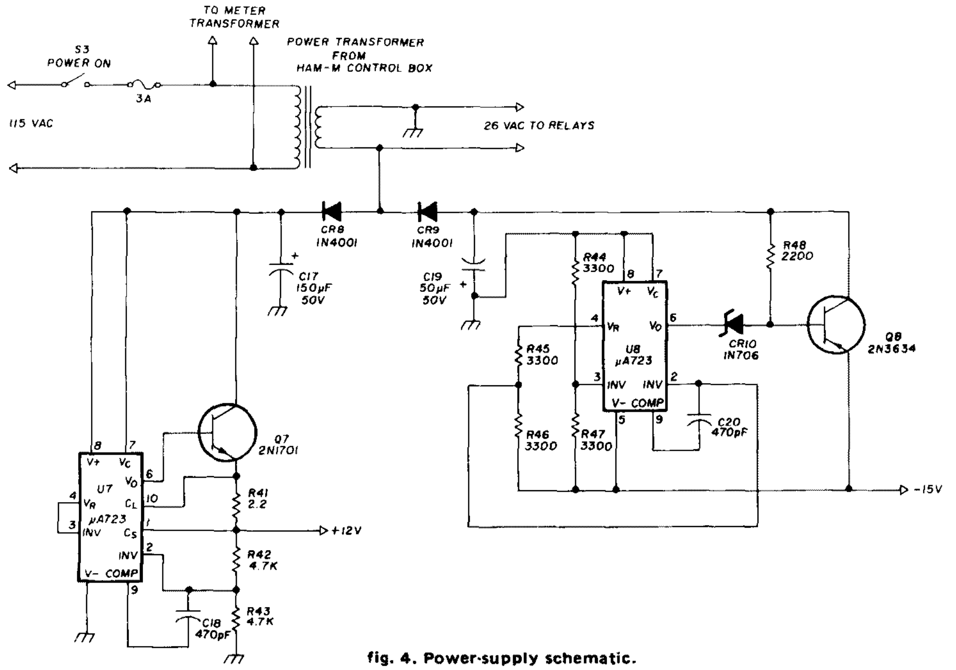


fig. 4. Power-supply schematic.

Position control R9 deserves some comment. I used a standard 270° rotation pot with a pointer knob. The panel was labeled S-W-N-E-S. If you can obtain a 360° rotation pot, the front panel could include a map to show the area to which your antenna is aimed.

LEDs 1 and 2 and switches 1 and 2 are needed only for testing and can be mounted inside the box. I also recommend that the connections between U3 and U5 and between U4 and U6 be jumper wires that can be easily removed for testing.

## testing and adjustment

This control circuit, if carefully built, can be connected to your Ham-M rotor without removing it from the tower for the following procedure.*

First check the control operation with the rotor disconnected by using a 500-ohm linear pot in place of the pot in the rotor. The meter should indicate the position of this temporary pot. The meter can be calibrated initially by adjusting R5 to give a full-scale meter reading when the temporary pot is turned fully to the right. When the temporary pot and the position control pot, R9, are in the same position, all the relays and LEDs (except of course the power indicator) should be off. As the pots are turned, the relays and LEDs should go through the sequence of turning the motor and brake on and off, as discussed previously.

When you're satisfied that the control circuit is working properly, set R17 and R20 for the widest dead band, remove the jumpers between the comparators and the timing circuit, then connect the rotor. The rotor can be operated by using the test switches, S1 and S2. Your antenna may require more brake delay than mine did. If so, increase timing capacitors C13 and C14 until the antenna comes to a smooth stop. Adjust the meter calibration if necessary. Check to see that the comparators are still functioning as indicated by LEDs 1 and 2.

Now, reconnect the comparators to the timers and check out the automatic positioning control. The final step is to adjust the dead band. Adjust R17 and R20 to make the dead band smaller, then rotate the antenna. When the dead band is too small, the antenna will coast through it before the brake is engaged. After the time delay, the motor will reverse only to coast through the dead band again. Increase the dead band until this searching action doesn't occur.

This automatic rotor control has proved satisfactory in two years of use at my station. A self-addressed stamped envelope would be appreciated with any correspondence relating to this article.

**ham radio**

*This design is based on a series-5 rotor. Series 1 and 2 rotors will require changes in the rotor.

# fine tuning the phased vertical array

Design and construction of a 2-element vertical in which phasing is stepped between zero and 180 degrees

Several years ago I described an antenna using two mobile whips mounted on the terrace railing of my apartment.[1] The antenna was a close-spaced array for 20 meters. The primary purpose of the design was to provide a means of reducing interference during reception. A secondary purpose was to provide some gain in line with the north-south pattern of the array. A simple switching arrangement provided a choice of zero degrees for no directivity, 135 degrees for a reversible cardioid pattern, and a spare position for "in between" phasing for experimentation.

While experimenting with this array over a two-year period, I made many observations, which convinced me (strictly from a reception standpoint), that it would be desirable to be able to shift the phasing quickly over a range between zero and 180 degrees in as many small steps as practical. This article describes a phase-shifting unit that accomplishes this objective.

## background

The popularity of the two-element phased array has been due to the work of Brown[2] and the charts he prepared (fig. 1) and, of course, to Kraus.[3] Since the publication of these articles many others have appeared in the literature. As background reading, I recommend a series of 12 articles on vertical antennas, which were published in *CQ* magazine.[4]

Brown's chart[2] was prepared under a given set of conditions that required the currents in each driven element to be of equal magnitude. This can be accomplished by power dividers and branching networks, several of which have been described in the amateur literature. The Wilkinson divider[5], the jeep-coil type, and the ohms-law type[6] are examples.

Most of the published amateur work on phased arrays has followed the rule of equal currents and has adhered pretty closely to the standard pattern structure and phasings as shown in **fig. 1**. Very few amateurs have ventured

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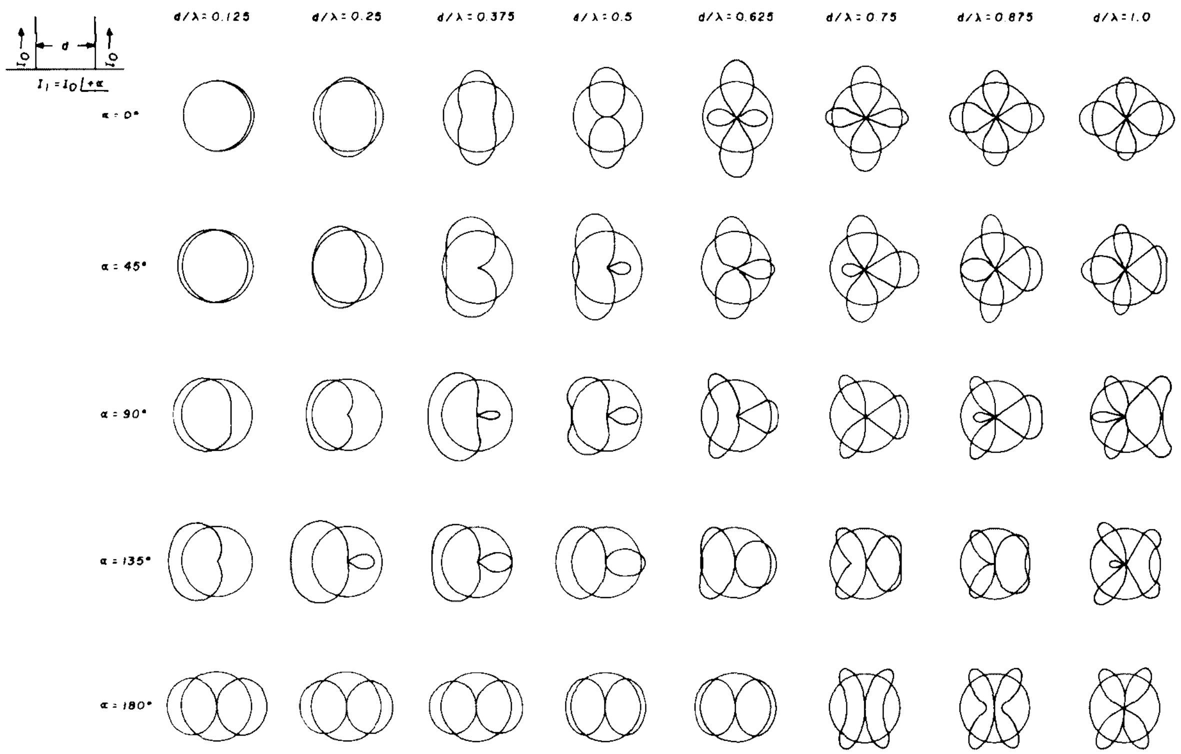


**fig. 1. Horizontal patterns for two vertical radiators fed with currents of equal magnitude (after Brown, reference 1). Circles indicate the relative field from one radiator alone with the same power input.**

the region of unequal currents and "in-between" phasing.

In an antenna system where the phasing will be varied from zero to 180 degrees, it's not a simple matter to keep the currents equal. In this case, the fact that the currents usually are not equal allows us to use those variations of the pattern from the norm created by a wide assortment of phase shifts.

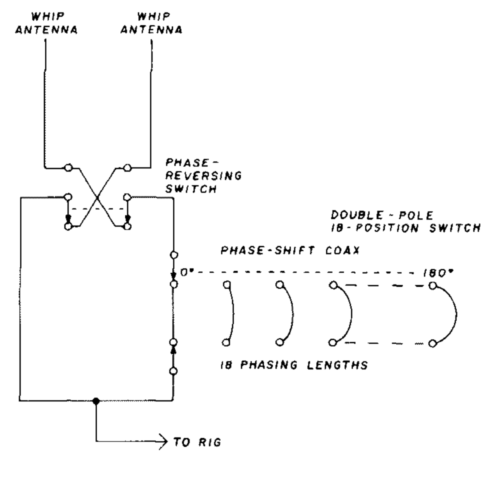


**fig. 2. A method for shifting phase in 10-degree steps. Eighteen separate pieces of coax cable would be required — not a very practical scheme.**

## null variations

For practical purposes the resultant patterns follow the classical patterns fairly well insofar as gain and directivity are concerned, but the main difference seems to be in the intensity and directivity of the nulls. Broadcast stations, for example, place their antennas so that the nulls avoid interference with other stations using the same frequency. Their patterns require a power division usually not equal to, and a phasing seldom a multiple of, 45 degrees.

The ease of reversing the directivity of a phased array quickly, with the flick of a switch, proved valuable in making observations that would have been otherwise impossible with a slow-moving rotary beam. All observations were made on the 20-meter band using two mobile whips spaced 12 feet (3.7m) apart in a north-south plane, mounted on the terrace railing of my apartment, which serves as a rather skimpy ground plane. The antennas are not vertical but extend out from the terrace at a 45-degree angle.

It was observed that the intensity of the null in the cardioid pattern obtained with approximately 135-degree phasing varied greatly as the vertical angle[7] of arrival responded to ionospheric changes and could be altered by a shift in phasing. This condition also existed

on the nulls in a figure-8 pattern obtained with 180-degree phasing.

Another observation made was the fact that there was an apparent shift in the azimuth of the deep null of the cardioid pattern, perhaps as much as 45 degrees, by small shifts in phasing. This phenomenon was also noticed in the nulls of the figure-8 pattern but to a smaller degree. This proved very helpful in eliminating interference on reception without materially affecting forward gain. Also, in the case of noise, a minimum could be obtained by proper phase selection; and in many cases the noise level could be reduced from an S-9 to an S-2 level.

The variable phase shift unit described here can be used with any 20-meter phased two-element array, regardless of spacing or polarization, and will allow you

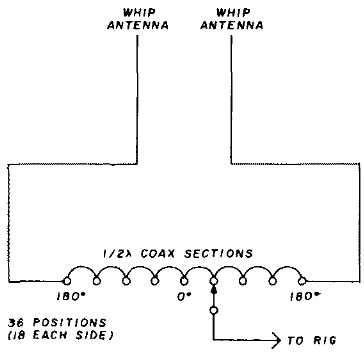


fig. 3. Simple and practical method for an antenna phasing system using a half-wavelength of coax cable with as many taps as desired.

to select the most suitable pattern and then proceed to "fine tune" the system to ionospheric conditions and interference, which are both usually in a state of flux.

## design

The original phase-shift unit I had been using with the two mobile whips provided a selection of four phasing positions, which were entirely inadequate for the "fine

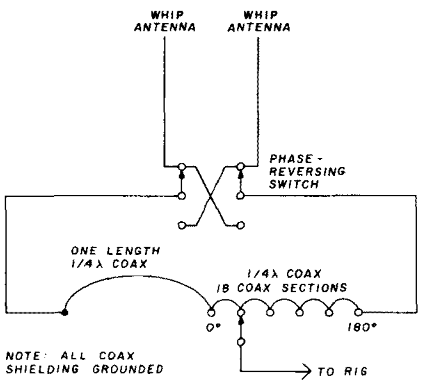


fig. 4. System as shown in fig. 3 but with only one side of cable tapped and a dpdt switch used for phase reversal.

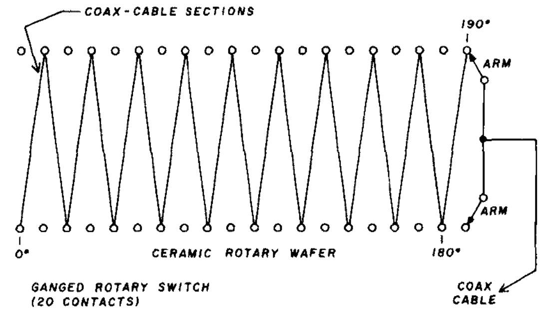


fig. 5. Method for connecting small-length coax sections to a 20-position rotary switch. The switch was modified so that the wafers were far enough apart to accommodate each coax-cable segment. The switch should be able to carry the rf current.

tuning" needed. I could have provided additional switching and shifted phase every 10 degrees, but I would have ended up with 18 separate rolled-up coils of coax, an unwieldly arrangement to say the least (**fig. 2**).

A quick solution appeared: why not just use one length of coax cable, 1/2-wavelength long, for the 180-degree phase shift and divide it into 18 sections so it could be tapped? Simple, but it won't work; the unused portion of the cable will act as an open-ended stub and detune the entire system.

A simple and practical method evolved in which there would be no unused section of cable. The entire half wavelength of coax would straddle the two antenna feedlines (**fig. 3**). At the centertap, the two antennas would be in phase (0 degrees) and at either end would be 180 degrees out of phase. It's apparent that as the tap is shifted, the amount of coax added to one feedline is automatically subtracted from the other. Of course, a further simplification is to tap one side of the line and use a dpdt lever switch for phase reversal (fig. 4).

The electrical length in inches* of a quarter-wavelength line (90 degrees phase shift) is determined by:

$$L = \frac{2952k}{F} \qquad (1)$$

$L$ = length (inches)
$k$ = coax-cable velocity factor
$F$ = frequency (MHz)

Inasmuch as coax cable varies, the line should be trimmed to frequency with the aid of a grid-dip meter. After the proper length is found, the number should be divided by 18 to give that many pieces of cable for 10-degree steps in phase shift. One additional piece of cable should be added to provide an overlap that may be needed for full 180-degree coverage because of differences in rotary switches and workmanship. For example, at a frequency of 14250 kHz you might grid-dip a quarter wave of coax and find it to be 135 inches (342.9cm). If so, each of the segments will be 7.5 inches

*For coax length in millimeters, the factor 2952 in **eq. 1** becomes 74980; for the length in centimeters it is 7498. **Editor**

(19cm) long. Adding one length will give 19 sections with a total length of 142.5 inches (363cm).

After the 19 sections of coax have been mounted on the switching assembly, the unit should be grid-dipped as a quarter wave and the frequency noted. This new frequency will be the guide for cutting a matching piece of coax to balance the system so the centertap position will be zero degrees.

A mechanical problem occurs in trying to cram 19

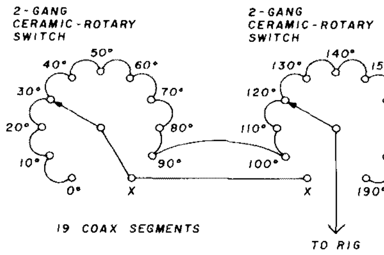


**fig. 6. Switch arrangement using two ganged sections. One phasing coax-cable segment is used as a link between each ganged section. Whichever switch is not in use should be set at contact marked with an X.**

short lengths of RG-58/U coax around a 20-position switch. This problem was solved by modifying a two-section rotary ceramic switch so that the wafers were placed just far enough apart to accommodate the length of coax segment. The coax was then placed lengthwise and attached to alternate contacts on each switch as shown in **fig.** 5 and using a common rotor for selection. The unused portion of the coax on the selector arm is short and of little consequence. The rotary switch, of course, should be of good quality and be able to carry the rf current. A word of caution: *do not* rotate the switch on transmit!

RG-58/U cable is used throughout the phasing unit because of its small size and ease of handling. Unfortunately, the use of this cable will limit the power to somewhat less than may be desired. I've been using 300 watts of key-down power output, but the coax should handle twice that power and even more. Of course, larger cable can be used as feedlines for the antennas.

The 18-segment number was chosen so that phasing could be divided into 10-degree steps, and the added segment to 190 degrees. This 10-degree number can be varied to conform to the amount of contacts available on your rotary switch. If you can accommodate more than 20 positions, fine. The more positions you have, the greater the selection of phasings and the shorter will be the coax sections. Conversely, fewer contacts mean longer segments and fewer phasing selections. I would limit the size of the segments to between 5 and 10 inches (13 and 26cm).

If you can't obtain a rotary switch with sufficient contacts, I suggest you use two ganged assemblies as shown in fig. 6. Mount the two separate ganged assemblies close enough together so that one phasing segment can be used as the link between them. Whichever switch is not in use should be set at contact X.

In preparing the coax for mounting on the switch, try to leave a minimum of exposed inner conductor. The shield should be wrapped together with a fine copper wire, then soldered (**fig. 7**). Also, the phasing unit internal wiring should be balanced right up to the fixed and variable phasing lines. Coax should be used in the internal wiring unless the leads are very short.

## matching impedances

Close-spaced elements will usually have a fairly low impedance because of the mutual impedance involved and will usually be in the neighborhood of about 15 ohms. An swr of even 3:1 can be tolerated unless the feedline is several wavelengths long. However, a simple linear matching transformer can be used consisting of 1/4 wavelength of two lengths of 52-ohm coax in parallel, grid-dipped to frequency. This section is used to match the 15-ohm impedance of the antennas to the 52-ohm feedline. This follows the formula for quarter-wave linear transformer design where the surge impedance of the linear transformer is equal to the square root of the product of the input and output impedances. The 26-ohm impedance of the parallel coax lines comes out pretty close to the desired value.

## checking and adjustment

Once the two antennas are connected together in the phasing unit, the output of the unit should be matched to the rig with an antenna tuner. The lead between the tuner and the phasing unit should be as short as possible and the tuner adjusted for minimum swr with the phasing set at 130 degrees. This setting will usually allow other phasing settings to be used *without having to* readjust the tuner (**fig. 8**).

This phasing unit can also be used on 10 and 15

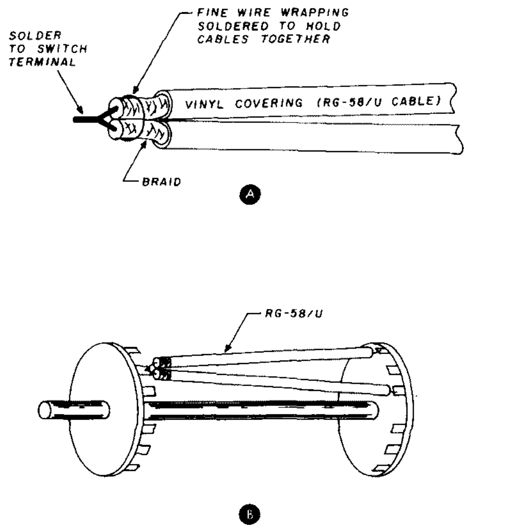


**fig. 7. Method for mounting coax cable to rotary-switch segments. Sketch A shows how to secure the cable ends. Sketch B is an example of the coax-cable segment in (A) mounted to a two-gang ceramic rotary switch.**

meters if desired without any changes. The calibration will be doubled on 10 meters to indicate 20 degrees per division and 15 degrees per division on 15 meters. A phasing unit can be designed for 80 meters and also used on 40 meters.

The effectiveness of the phasing unit can be checked and calibrated in the receive position by using a steady signal source such as a grid-dip meter placed 30 or 40 feet (9 or 12m) away from the array, first in the position where the cardioid null falls, then where the null of either side of the figure-8 falls. The phasing setting that provides the best front-back ratio for the cardioid pattern should be designated as 135 degrees (130 degrees will do); and for the best broadside nulls, 180 degrees. The calibration can then be verified by tests with local stations in known directions. The other calibration points can be determined by interpolation.

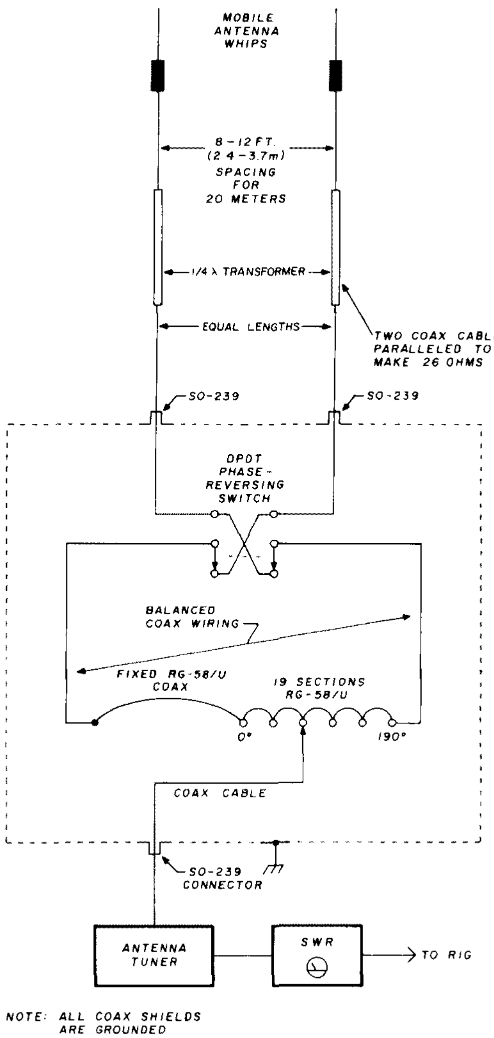


**fig. 8. Final schematic diagram of phasing unit shown connected to antennas, tuner and swr indicator.**

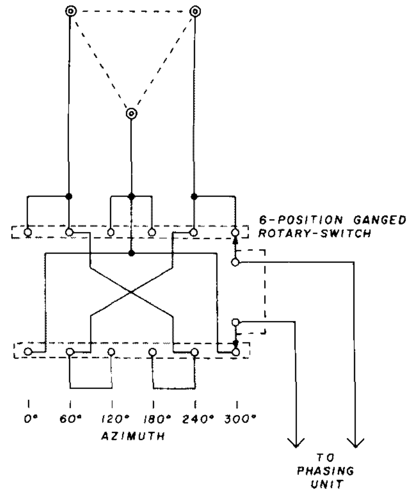


**fig. 9. Arrangement of three vertical antennas in a triangle for full directional coverage.**

In operation, when in contact with another station, the phasing control is usually set at 135 degrees with the reversing switch set for strongest reception. If interference is present, an attempt is then made to reduce it by adjusting the phasing control. Sometimes reversing the array will reduce interference more than the wanted signal. If this occurs, simply flip the reversing switch back to normal for transmitting. This situation will occur often when there is a high noise level, which can be reduced considerably by reversing the cardioid pattern. The figure-8 pattern is normally used when the interference or noise level is broadside to the array.

## concluding remarks

The flexibilty of this phasing system can be rewarding once the user becomes familiar with its capabilities. For those with space for three close-spaced verticals in a triangular arrangement, this system can be used with any two of the verticals. The advantage will be complete coverage in azimuth in six steps. The switching arrangement in **fig. 9** should do the trick.

I welcome comments from those who have built the phasing unit described here and have obtained experience with its use on the amateur bands.

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**ham radio**

# a comparison of vhf mobile antennas

## Mobile antenna gains depend not only on antenna type but — deflection due to vehicle speed

The issue of antenna gains, especially at vhf, seems to be shrouded in some mystery. At times, the gain figures of antennas appear to be determined more by the sales department than the engineering department. As part of my job I had the occasion to test a number of vhf mobile antennas to determine their in-use performance. It is hoped the results will prove useful to the amateur fraternity.

### gain references

Usually the gain of a base station antenna is referenced to a half-wavelength dipole and the gain of a mobile antenna to a quarter-wavelength monopole. The Electronic Industries Association (EIA) has previously established standards for both base and mobile antennas. These standards are EIA-RS-329, part 1, for base antennas and EIA-RS-329, part 2, for mobile antennas. Commercial antenna manufacturers now rate the gains of their land-mobile base antennas per EIA-RS-329 although few, if any, rate their mobile antennas by the same standard.

### antenna gains

Mobile antennas are usually tested on a 56-inch (22cm) ground screen or on a standard automobile. In the testing presented, all the antennas were mounted on the rear cowl (passenger side) of a Dodge police cruiser. All the antennas were adjusted for a vswr of less than 1.5:1. Each test consisted of transmissions from the mobile to a base receiver using a DB-224 antenna at 175 feet (53m). Signals were then recorded on a calibrated limiter meter using step attenuators as required. The tests were performed at various locations at distances of 1 to 20 miles (1.6 km to 32 km) from the base station. At every location two tests on each antenna were performed, one with the vehicle pointed toward the base station and the other with the vehicle pointed away from the base station. **Table 1** shows the average gains referenced to a quarter-wavelength whip with the vehicle stationary.

Although above data was taken at 155.5 MHz, the gain figures are equally applicable at 146 MHz. The gains are somewhat higher than might be expected due to the effects of the roof of the vehicle. The roof shields the quarter-wavelength antenna more than the five-eight wavelength antenna when the auto is pointed toward the base station. In all tests, however, the relative gains were in the same order as the final averages.

Since we have the antenna on an automobile, gain with the vehicle stationary is only part of the story. Tests were also performed to see how much degradation was suffered at high vehicle speeds. Since testing was being done for a police agency, high speed meant 100 mph (160kh/h)! **Table 2** shows the antenna deflection at 100 mph.

As mentioned previously, the test frequency was 155.5 MHz; at 146.5 MHz the antennas are longer, 3 inches (1cm) for five-eighth wavelength antennas so you would anticipate deflection to be somewhat greater.

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**table 1. Stationary antenna gains. Antenna gains with reference to the 1/4-wavelength whip. Only two measurements were performed on the DB-702. The ASP-800 is a heavy duty version of the ASP-177 with approximately 0.5 dB more gain.**

| antennas | stationary average gain |
|---|---|
| 0.046 inch (1mm) stainless 1/4-wavelength whip | 0 dB |
| 0.100 inch (2.5mm) Larson 1/4-wavelength whip | — |
| Antennas Specialists 800 5/8 wavelength 0.125 inch (3mm) whip | 3.1 dB |
| Antennas Specialists 800 5/8-wavelength without spring | 3.1 dB |
| DB Products 702 5/8-wavelength 0.100 inch (2.5mm) whip | 2.2 dB |
| Larson 150 5/8-wavelength 0.100 inch (2.5mm) whip | 3.6 dB |
| Larson 150 5/8-wavelength 0.125 inch (3mm) whip | 3.6 dB |

Subsequent tests with an antenna cut for 146 MHz indicates approximately the same antenna deflection at vehicle speeds of 70 mph (113km/h).

## high speed gain

Tests were then run to check the effect of antenna deflection on performance. Table 3 shows both the high-speed gain reduction and the gain at high speed with reference to a quarter-wavelength whip. At high speeds the gain degradation is both apparent and dramatic. The five-eighth wavelength antennas with 0.100-inch (2.5mm) whips suffer much greater degradation than the five-eighth wavelength antennas with 0.125-inch (3mm) whips or the quarter-wavelength antennas.

Degradation was not caused by detuning of the antenna or transmitter, since there was only a slight increase in vswr when the antennas were deflected. Although the 0.046-inch (1mm) quarter-wavelength whip and the ASP-800 deflected approximately equal amounts (45 degrees and 50 degrees respectively), the ASP-800 had significantly greater degradation. This greater degradation is caused by the narrower beamwidth of the gain antennas being bent up above the horizon more than the wide beamwidth of the quarter-wavelength antenna.

**table 2. Antenna deflection at high vehicle speed. Speed of approximately 70 mph (113km/h) produced above deflection with antenna cut for 146 MHz.**

| antenna type | deflection |
|---|---|
| 0.046 inch (1mm) stainless 1/4-wavelength whip | 45 degrees |
| Larson 0.100 inch (2.5mm) stainless 1/4-wavelength whip | 20 degrees |
| Antennas Specialists 800 5/8-wavelength 0.125 inch (3mm) whip | 50 degrees |
| Antennas Specialists 800 5/8-wavelength whip without spring | 30 degrees |
| DB Products 702 5/8-wavelength 0.100 inch (2.5mm) whip | 80 degrees |
| Larson 150 5/8-wavelength 0.100 inch (2.5mm) whip | 80 degrees |
| Larson 150 5/8-wavelength 0.125 inch (3mm) whip | 30 degrees |

This was verified in testing of uhf antennas when a 30-degree deflection produced a degradation of 5.5 dB for an antenna that used two vertically phased five-eighth wavelength elements. This compares to a 30-degree deflection producing a 3.0 dB degradation for a single five-eighth wavelength element.

It is apparent that for maximum performance it is necessary to maintain the antenna vertical and still retain immunity to damage from overhead objects. Knowing that the antenna must be kept vertical, placement on the vehicle becomes important. Tests by Antenna Specialists[1] show a 2.5 dB increase in signal when a vhf quarter-wavelength antenna is moved from the trunk lip to the center of the roof. The five-eighth wavelength antenna shows a 1 dB increase when moved from the trunk lip to the center roof.

Since most of my driving is within 10 miles (16km) of the local repeater, a quarter-wavelength whip on the roof is adequate. When traveling out of town a five-eighth wavelength antenna (with a 0.125 inch (3mm) whip) is substituted on the same mount. The quarter-wavelength antenna offers an added advantage around town. Because the quarter-wavelength whip doesn't resemble a CB antenna, the radio is somewhat more immune to theft since stolen amateur equipment doesn't have as large a market as do stolen CB radios. In some areas this aspect may outweigh the performance factor.

**table 3. Antenna gain reduction at high speed and the resultant gain with reference to a 1/4-wavelength whip at high speed. The 5/8-wavelength antennas without the 0.125 inch (3mm) whip actually have a loss with respect to the 1/4-wavelength whip at high speed.**

| antenna | average gain reduction at high speed | average gain at high speed |
|---|---|---|
| 0.046 inch (1mm) stainless 1/4-wavelength whip | 1.6 dB | 0 dB |
| 0.100 inch (2.5mm) Larson 1/4-wavelength whip | 0.9 dB | 0.7 dB |
| Antennas Specialists 800 5/8-wavelength 0.125 inch (3mm) whip | 4.1 dB | 0.6 dB |
| Antennas Specialists 800 5/8-wavelength without spring 0.125 inch (3mm) whip | 3.2 dB | 1.5 dB |
| DB Products 702 0.100 inch (2.5mm) whip | 7.6 dB | − 3.8 dB |
| Larson 150 5/8-wavelength 0.100 inch (2.5mm) whip | 7.3 dB | − 2.1 dB |
| Larson 150 with 0.125 inch (3mm) whip | 3.0 dB | 2.2 dB |

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**ham radio**

# high performance antenna for 80 meters

## Simple construction and high performance are two features of this 80-meter phased array

While tuning around the different amateur bands, you can find that the antennas in use fall into a familiar pattern. On 10, 15, and 20 meters, most people are using a Yagi or quad; on 40 meters, the majority are using a vertical or dipole. However, on 80 meters, a wide variety of antennas is encountered, the inverted Vee being the most common. The larger stations do have a wide range of antennas; Rhombics, long wires, long horizontal vees, phased verticals, big loops, collinears, Yagis, quads, occasionally even a Beverage.

At my location in Tennessee, close to the center of the United States, any antenna I install for 80 meters seems to work well for stateside contacts. However, I generally run into trouble busting through the "copper curtain" of the East and West Coasts to DX contacts. This presents no big problem for ordinary hamming and rag chewing. When I decided to climb on the bandwagon and work for 5BDXCC though, I learned quickly that it was going to be a struggle unless I installed an antenna a little better than an ordinary vertical or dipole. I have a conventional horizontal dipole, about 60 feet (18m) above ground, that has been up for several years and works fairly well, but not good enough to stand up under competition for the long-haul contacts or in the pile-ups.

### antenna comparison

Not being hampered by a restricted house lot or lack of real estate, I was able to leave up the old dipole and still install other antennas at the same time. The dipole was used as my standard of comparison. As each new type of antenna was erected, I would compare the performance of the new one with the old dipole. In a fairly reasonable period of time this procedure would allow me to determine whether the new antenna was a bomb or a dud.

I've learned that judging the performance of a single antenna in just a couple of contacts, or over a short period of time without some means of comparison, is inconclusive. When band conditions are good, even a short length of barbed wire fencing will give you a few contacts. Conversely, if you check an outstanding antenna system when conditions are poor, you can often be erroneously led to believe that the antenna was a complete waste of time. So over a period of several years, I've spent many weekends installing new antennas for 80 meters and checking them against my old dipole.

The relative performance of the more common antennas surprised me in that several *seemed* to perform beautifully. It wasn't until I compared them to my dipole that I learned they weren't all-round performers, or didn't meet the claimed performance standards.

During the last five years, I've installed and tested verticals, phased verticals, inverted-vees at different heights and configurations, multi-element inverted-vees, horizontal vees up to 650 feet (200m) long, long wires several wavelengths long, loaded rotatable dipoles, discones, sloping dipoles (both single and multi-element and with reflectors) Zepps, loops, shortened quads, collinears and even a small rhombic. I haven't yet gotten around to installing a good full-size rhombic, but this is on the program. Several antennas did outperform the dipole, but their limitations and the small improvement in performance didn't always justify or warrant the extra hardware, weight, and real estate that was involved.

In the course of trying for a better antenna, I attempted to feed in phase a second dipole, identical to the old one, parallel to it and at the same height. This seemed to offer very little improvement. In fact, any improvement of receive or transmit signal strengths were practically indiscernible, in any direction or over any distance.

### phased array

Before tearing down this system, I decided to see what would happen if I fed the two antennas 90 degrees out of phase, by very simply inserting a 42 foot (13m) length of coax in the feedline to one of the dipoles. The results were startling in that performance came close to equalling the best multi-element or long wire antennas I had ever used.

I found that by spacing the two dipoles parallel to one another, about 120 feet (36.5m) apart, and at approximately the same height, I was getting about 3-5 dB of forward gain; front-to-back ratio was about 12 dB. Even with some forward gain in either of two directions, my side lobes are broad enough to permit operation off the side with a respectable signal. My two dipoles are

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oriented north-south, with the gain and front-to-back to the east and west. I could improve the performance by varying the spacing of the two antennas, but the convenience of using trees already present, and the fact that I like my comparatively omnidirectional pattern away from the prime direction discourages my changing the existing spacing. By running the feedlines from the two dipoles to a coax switch, and inserting a 42-foot (12.8m) length of RG-8/U into the feedline of either the east or west antenna, I can switch the directivity to either the east or west (fig. 1). Signals from north and south don't seem to be affected very much.

After this antenna installation was made, my total of 50 countries on 80 meters jumped to 140 countries in the next three months. This, without too much effort or more than a casual attempt at DXing on 80-meter ssb and CW.

The only non-conventional aspect of my dipoles is

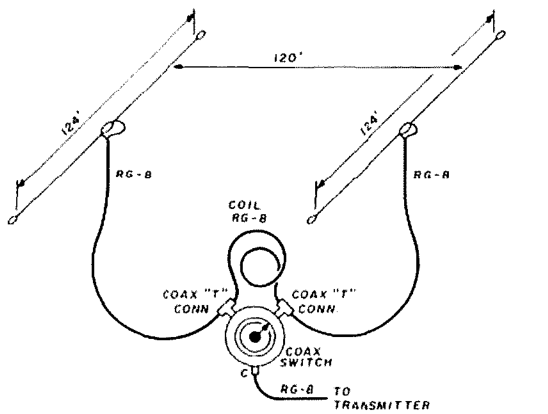


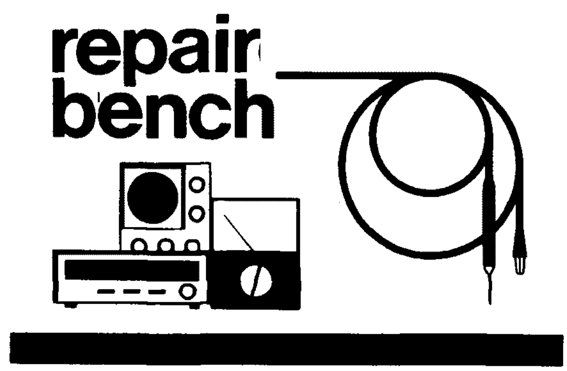

**fig. 1. The antennas are fed through identical lengths of coax. By tuning each antenna individually to the same frequency, it is possible to switch directions without retuning the amplifier. The coax switch cannot be self-grounding.**

the feedline. Instead of a normal coax feed, I use a standard quarter-wave unbalanced-to-balanced transformer (balun) feed system. The balun seems to improve the performance of the dipoles. The only caution that must be exercised is the exact match of the two feedlines and the two antennas. Otherwise, the transmitter or linear will have to be retuned after changing the direction. I found that antennas of the same length were not entirely satisfactory. The second antenna was affected by several nearby trees, the house, and a slight difference in height. By changing the length to get it to load exactly like the first dipole, I can switch directions without retuning.

The array is broadband enough to permit operation on both the ssb and CW portions of the band without exceeding a swr of about 2.8:1 at the ends of the band. It even permits operation on the Army MARS frequency above the high end of the 75-meter phone band.

For a simple, easy to install and tune, and inexpensive arrangement that has a minimum of hardware, it's the best all-around antenna I've ever used on 80 meters.

**ham radio**




# Bob Stein, W6NBI

## how to use the slotted line for transmission-line measurements

In a previous article,[1] several applications of the swr indicator (such as the Hewlett-Packard Model 415B) were described, *excluding* its use in conjunction with a slotted line. Since it was for that specific purpose that the swr indicator was developed, this article will actually be a continuation of the previous one as well as one which will discuss the use of coaxial slotted lines to measure swr, impedance, and transmission-line loss. The frequency range over which coaxial slotted lines are used is typically 400 to 5000 MHz, although it is possible, under certain conditions, to go as low as 100 MHz.

Since the purpose of this series of articles is to popularize the use of test equipment which is often available on the surplus market, let's begin by enumerating some of the coaxial slotted lines which you may run across. Hewlett-Packard Models 805A and 805C and General Microwave Type N200 are slab-type lines, which will be discussed shortly. Conventional coaxial lines, such as the GenRad (formerly General Radio) types 874-LBA and 874-LBB, are also available. Military types IM-24/U, IM-25/U, IM-92/U, and others are usually nothing more than the aforementioned commercial models which have been assigned military nomenclature. One exception to this is the TS-56A/AP, which appears to have been designed originally as military equipment.

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### principles of the slotted line

The principles governing the operation of the slotted line are those which apply to all rf transmission lines, and which will not be repeated here except as they apply directly to using the line. If we consider a section of transmission line which is fed from an rf source and is terminated in its characteristic impedance, $Z_o$, we know that there will be no reflection from the terminating load and therefore there will be no standing waves on the line. On the other hand, if the line is terminated by an impedance which is different from $Z_o$, a standing-wave pattern will result, as shown in **fig. 1**.

If the transmission-line section is coaxial, the voltage standing wave will exist as a potential difference or field between the inner and outer conductors. If we probe this field along the length of the line with a detector, and are careful not to disturb the field excessively with the probe, the detector will provide an output voltage which varies as the field intensity or voltage amplitude.

Since adjacent voltage minima or maxima along the line are always a half wavelength apart, the wavelength of the rf source can be determined by measuring the actual distance between adjacent minima or maxima. Furthermore, the detector is able to provide relative amplitudes of the voltage maxima, $e_{max}$, and minima, $e_{min}$, enabling us to determine the voltage standing-wave ratio from the following equations:

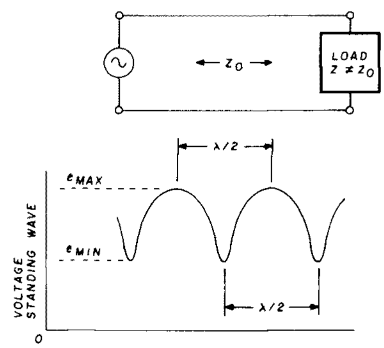



**fig. 1. Standing wave on a transmission line terminated by a load impedance which is not equal to its characteristic impedance.**

$$swr = \frac{e_{max}}{e_{min}}$$

$$swr\ (dB) = 20 \log \frac{e_{max}}{e_{min}}$$

Knowing the wavelength and the swr, it is also possible to determine the impedance of the load by means of transmission-line relationships. This will be explained later.

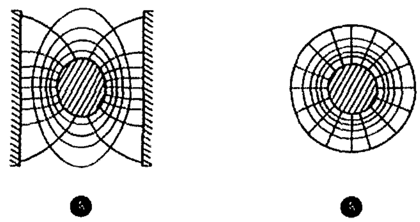


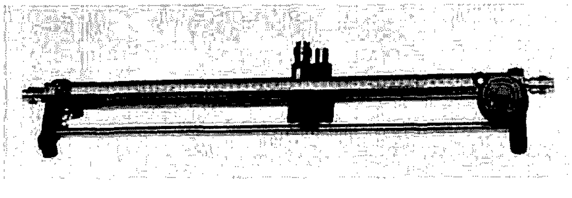
fig. 2. Cross-sections of the slab-type (A) and coaxial (B) slotted lines. The radial lines represent the electric field, the concentric lines the magnetic field. Courtesy Hewlett-Packard Company.

The conventional coaxial slotted line, such as the GenRad Type 874-LBB shown in the photograph, consists of a length of precision 50-ohm coaxial line with a narrow slot cut longitudinally along the outer conductor. A probe, mounted on a carriage which is movable along the length of the slot, extends into the line to sample the rf field. Connected to the probe is a microwave diode; a tuning probe or stub tuner (not shown in the photograph), also connected to the probe and diode assembly, tunes the structure for maximum sensitivity. A scale, graduated in centimeters and millimeters, is attached to the frame so that a pointer on the carriage can be used to measure distances along the line.

A somewhat different line configuration is the slab line used by Hewlett-Packard for their model 805 series and by General Microwave for the Type N200. As shown in the photograph of the Hewlett-Packard Model 805C, the probe carriage is mounted on a box-like structure which is actually two parallel conducting semi-planes separated by about 0.8 inch (20mm) and between which is the center conductor. Fig. 2 shows cross-sections of the slab line and the conventional coaxial line. The equipotential field lines show that each of the semi-planes of the slab line is equivalent to one-half the outer conductor of the coaxial section. Precision machining results in a 50-ohm characteristic impedance and, according to Hewlett-Packard, the space between the semi-planes is equivalent to a slot width of less than 0.002 radian in a coaxial line.

The probe carriage contains a microwave diode detector, the probe, and a probe tuner. A scale, calibrated in centimeters and millimeters, is mounted on the frame and is used in conjunction with a vernier scale on the carriage to permit wavelength resolution to within 0.1 millimeter (about 0.004 inch).

In both types of slotted lines, the detector is usually a type 1N21 or 1N23 microwave diode, operating in its square-law region. Bolometer elements, such as a Narda Type N821, a PRD Type 610-A, or a selected 10-milliampere instrument fuse, may also be used. Bolometers require less attention relative to square-law operation, but are less sensitive than a diode detector.



The GenRad (formerly General Radio) Type 874-LLB Slotted Line. An adjustable stub or probe tuner is usually inserted into the left-hand connector on the movable carriage. Photo courtesy GenRad.

For impedance measurements, a precision short must be connected to one end of the slotted line. Because the GenRad line uses hermaphrodite GR874 connectors, the GR Type 874-WN3 Short-Circuit Termination will fit either end. The Hewlett-Packard and General Microwave slotted lines are equipped with type-N connectors, male at one end of the line and female at the other. These lines are supplied with precision male and female shorting terminations.

The low-frequency limit of the slotted line is a direct function of its usable length, that is, the distance over which the probe carriage can travel. The Hewlett-Packard Model 805A and 805C frequency range is specified as 500 to 4000 MHz, although these lines are usable down to about 400 MHz, since the probe travel is approximately 36.5 centimeters (one-half wavelength at 410 MHz). The GenRad Type 874-LBB is rated from 300 to 8500 MHz, based on its probe travel of 50 centimeters. The earlier Type 874-LBA has a specified upper frequency limit of 5000 MHz.

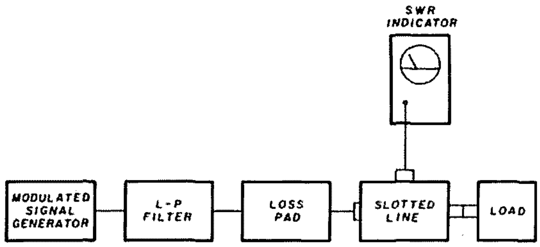



fig. 3. Equipment set-up for measuring swr, using an swr indicator for direct measurement readings.

These slotted lines can be used at frequencies below their specified minima, but such measurements may require the use of additional air-dielectric transmission-line sections or line stretchers. Such applications are beyond the scope of this article, but are covered in reference 2.

## connecting the slotted line

The equipment set-up for using the slotted line is shown in fig. 3. The swr indicator was discussed in detail in reference 1. The signal generator must be amplitude modulated with a sine or square wave at the same frequency to which the swr indicator is tuned – usually 1000 Hz. Although an output of 1 milliwatt into 50 ohms is recommended for measurement of high standing-wave ratios, an output of 0.2 milliwatt (100 millivolts across 50 ohms) will generally suffice. The generator must supply a constant output which has minimum harmonic distortion and low incidental fm. The modulating voltage must also be stable to minimize measurement errors.

A lowpass filter, connected to the output of the signal generator, is desirable although not absolutely

necessary. Severe harmonic distortion in the signal-generator output can result in erroneous measurements, but these are probably of minor importance for amateur use. A loss pad of at least 6 dB should be used to minimize loading effects of the test set-up on the signal generator.

All of the connections to the input of the slotted line can be made with interconnecting coaxial cables, but the load must be connected directly to the slotted-line connector. In the case of lines having a male connector at one end and a female at the other, connect the load to the end which will permit a direct connection or, only if absolutely necessary, one made with the fewest number of adapters. Otherwise, the swr of the adapter(s) will introduce errors into your readings.

The measurement procedures which follow are, of necessity, abbreviated and generalized so as to be applicable to most slotted lines. That is not to say that they cannot be applied directly to whatever line you may be using, for in fact, they can. But there are much more detailed instructions in the manufacturers' manuals, and I strongly recommend that either the applicable manual be obtained or that one of those specified in the references at the end of this article be purchased.

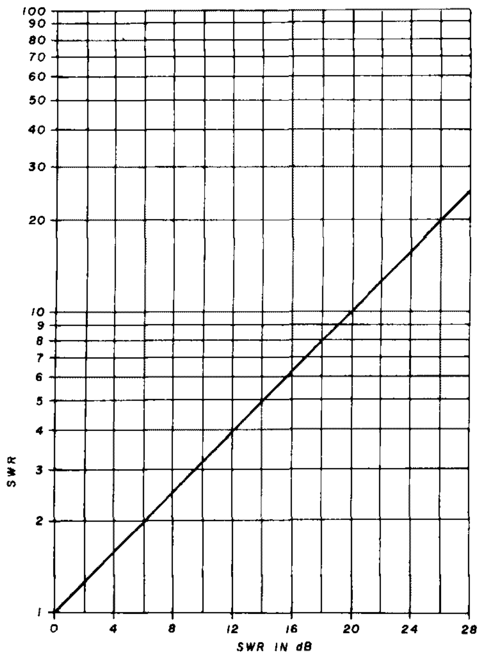



**fig. 4. Swr vs swr in dB.**

## measuring swr less than 10:1

**Direct method.** The direct method of measuring swr is that which is most often used, although the alternate attenuator method will also be described. The procedure for direct measurement of swr is as follows:

**1.** Connect the equipment as shown in **fig. 3**, except that the load is left unconnected; allow sufficient warm-up time for the signal generator to become stable.

**2.** On the swr indicator, set the *gain* control fully counterclockwise and the *range* switch to **50**.

**3.** Adjust the probe depth and tuning for a reading on the swr indicator, changing the *range* switch position clockwise, if necessary. If two peaks are obtained as the probe circuit is tuned, use that which results in the higher reading.

**4.** Connect the load to the slotted line and move the carriage along the line to find a voltage minimum.

**5.** Retune the probe circuit for a maximum meter reading. Then reduce the probe depth and retune the probe circuit (these adjustments interact) until a stable reading is obtained on the swr indicator with its *range* switch set to **50** or **60**. It may also be necessary to adjust the signal-generator output. The main object, however, is to achieve a stable meter indication consistent with *minimum probe insertion*.

**6.** Move the carriage along the line to find a voltage maximum, and adjust the swr indicator *gain* control and *range* switch to obtain a meter reading of *1* on the swr scale.

**7.** Move the carriage to obtain a minimum meter reading, without changing any other adjustments or controls on the slotted line or on the swr indicator.

**8.** Read the standing-wave ratio directly from the swr scale on the meter. If the swr is greater than **3**, switch to the next lower (counterclockwise) position of the *range* switch.

**9.** If the swr is less than **1.3**, a more accurate indication can be achieved by setting swr indicator *meter scale* switch to *expand*, and repeating steps **6** and **7**. In this case, the standing-wave ratio is read from the *expanded swr* scale on the meter.

**Attenuator method.** The attenuator method of measuring swr eliminates any error which may be introduced by deviation of the detector from true square-law response. It can also be used if an swr indicator is not available or if the signal source is unmodulated.

If the attenuator method is employed in conjunction with a *modulated* signal generator and an swr indicator, the test set-up of **fig. 3** is applicable, except that a precision variable attenuator is used in place of the loss pad. The measurement procedure is as follows:

**1.** Perform steps **1** through **7**, described under the direct measurement method, with the variable attenuator set to provide at least **6** dB of attenuation.

**2.** Adjust the *gain* control and the *range* switch on the swr indicator for a meter reading of **0** dB with the *range* switch set to **50** or **60**.

**3.** Move the carriage along the line to find a voltage maximum; do not readjust the swr indicator controls.

**4.** Increase the variable attenuator setting until the swr indicator again reads **0** dB, or as close to it as can be obtained on-scale.

**5.** The swr, in dB, is equal to the variable attenuator

setting plus the meter reading (also in dB), minus the variable attenuator setting used in **step 1**.

**6.** Swr in dB can be converted to swr by reading the value from the scale on the meter which corresponds to the dB value of swr. This relationship is also plotted in **fig. 4**.

If an swr indicator is not available, or if the signal generator is *unmodulated*, connect the equipment as shown in **fig. 5**, leaving the load initially disconnected. Then proceed as follows:

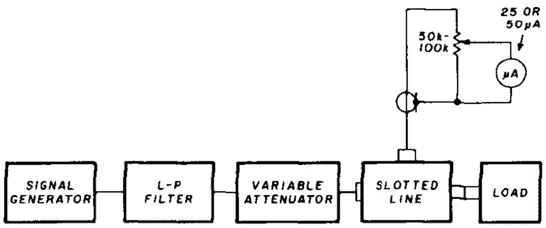



**fig. 5. Equipment set-up for measuring swr, using a microammeter for an indicator. The signal generator does not have to be modulated in this arrangement.**

**1.** Set the meter potentiometer to about mid-range and the variable attenuator to provide at least **6** dB of attenuation.

**2.** Adjust the probe depth and probe tuning for a reading on the meter, readjusting the meter pot if necessary.

**3.** Adjust the carriage position for maximum meter reading and retune the probe circuit for maximum. If two peaks are obtained as the probe circuit is tuned, use that peak which results in the higher current.

**4.** Connect the load to the slotted line.

**5.** Move the carriage along the line again to obtain maximum current.

**6.** Readjust the probe depth and tuning so that, with the meter pot set for maximum sensitivity, the meter current does not exceed **50** microamperes.

**7.** Move the carriage to obtain minimum current and note the meter reading.

**8.** Reset the carriage for maximum current, and increase the setting of the variable attenuator until the current is the same as that recorded in step **7**.

**9.** The swr, in dB, is equal to the variable attenuator setting minus the setting used in step **1**. Swr in dB can be converted to swr by the following expression, or from the curve of **fig. 4**.

$$swr = antilog \frac{swr\ in\ dB}{20}$$

## measuring swr greater than 10:1

It is extremely unlikely that any amateur will be interested in *accurate* readings if the swr of the device under test is greater than 10:1. The techniques for making such measurements are explained in detail in references 2 and 3, and do not warrant coverage here.

The load impedance on a transmission line can be calculated from a knowledge of the swr present on the line and the position of a voltage minimum with respect to the load. Although this seems complicated, the physical measurements using a slotted line are quite simple. Other than the equipment required to measure swr, a precision (low-inductance) shorting termination is needed.

The measurement procedures involve the precise location of adjacent voltage minima along the line. When the swr is low, the exact location of a minimum is difficult to establish because the minimum-voltage region is quite broad. Improved accuracy in locating the minimum may be achieved by averaging the carriage positions which provide equal voltages on each side of the voltage minimum, as follows.

**1.** *Mentally establish a convenient reference* which is approximately the mid-point between the maximum and minimum readings obtained on the swr indicator (or on the microammeter, if the attenuator method of swr measurement is used).

**2.** Move the carriage to one side of the minimum until the reference level is obtained on the indicating device. Note the carriage position in millimeters on the slotted-line scale.

**3.** Repeat **step 2**, moving the carriage to the other side of the minimum.

**4.** Average the positions obtained in **steps 2** and **3** by adding the two position readings and dividing by two. This is the position of minimum voltage.

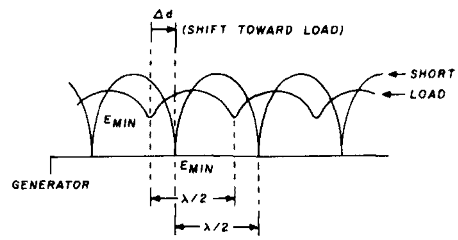



**fig. 6. Standing-wave patterns on a slotted line with a capacitive load and with a shorting termination. Courtesy Hewlett-Packard Company.**

The following procedures are used to measure the impedance of any load connected to the slotted line.

**1.** Measure and record the swr of the load, using any of the methods described previously.

**2.** Note the exact carriage positions in millimeters (from the scale) for two adjacent voltage minima. Record the difference between the two positions, which is equal to $\lambda/2$, as shown in **fig. 6**. Also record the position of one of the minima.

**3.** Replace the load with a precision short.

**4.** Move the carriage to find a new minimum which is closest to the one recorded in **step 2**.

5. Record the shift, $\Delta d$, between the positions recorded in steps **2** and **4** and note whether the shift is toward the generator or the load.

**6.** Calculate the electrical length, $\theta$, in degrees, of $\Delta d$ from the following expression

$$\theta = \frac{180(\Delta d)}{\lambda/2}$$

If the minimum shifted toward the load in step **5**, $\theta$ is considered positive; if the minimum shifted toward the generator, $\theta$ is considered negative.

**7.** Determine the impedance from the following equation (where $Z_o$ equals 50 ohms, the characteristic impedance of the system,

$$Z = Z_o \left[\frac{1 - j(swr)(\tan\theta)}{(swr) - j(\tan\theta)}\right]$$

or by means of a Smith chart, as explained below.

## determining impedance from a Smith chart

Solving the equation in step **7** above involves several rectangular-to-polar and inverse conversions, and is tedious, even with a scientific calculator which incorporates those conversion functions. Fortunately, the ubiquitous Smith chart provides a quick and simple solution to the problem. After proceeding through steps 1 through 6 under impedance measurements, continue as follows:

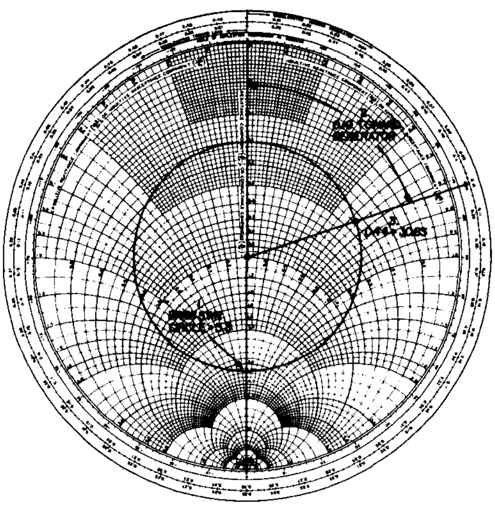


**fig. 7. Example of determining impedance by use of a Smith chart. Courtesy Hewlett-Packard Company.**

1. Convert $\Delta d$ to wavelength ($\Delta\lambda$) by means of the expression

$$\Delta\lambda = \frac{\Delta d}{\lambda}$$

where $\lambda$ is twice the value of $\lambda/2$ recorded in step **2** of the impedance measurement procedures.

**2.** Draw a circle, on the Smith chart, whose center is at the origin (1.0) and whose radius is equal to the measured swr.

3. Along the periphery of the Smith chart, mark a point equal to $\Delta\lambda$, the shift in wavelength, either toward the generator or toward the load, as applicable.

**4.** Draw a radius line from the origin to the point established in step 3.

5. Read the normalized impedance at the intersection of the radius line and the swr circle.

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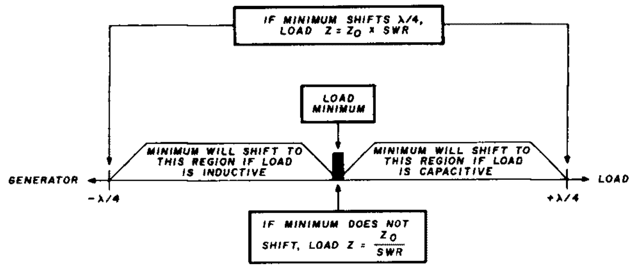



**fig. 8. Summary of rules for impedance measurements. Courtesy Hewlett-Packard Company.**

6. Multiply the normalized impedance by $Z_o$ (50 ohms) to convert to the actual impedance.

As an example, assume the following values were obtained by measurement:

$swr$ = 3.3
$\lambda/2$ = 150 mm, with one minimum at 220 mm
$\Delta d$ = 30 mm, toward generator

a. Calculate $\Delta\lambda$

$$\Delta\lambda = \frac{\Delta d}{\lambda} = \frac{30}{2(150)} = 0.10\lambda$$

b. Draw an swr circle with a radius equal to 3.3, as shown in **fig. 7**.

c. Draw a radius line for a shift of $0.10\lambda$ toward the generator.

d. **Fig. 7** shows the radius line and circle intersecting at $0.44 + j0.63$, which is the normalized series impedance of the load.

e. Multiply the normalized impedance by 50, giving the actual impedance as $22 + j31.5$ ohms.

## rules of thumb for impedance measurements

Some rules of thumb that are helpful when making slotted-line measurements are:*

a. The shift in the minimum when the load is shorted is never more than ± one-quarter wavelength.

b. If shorting the load causes the minimum to move toward the load, the load impedance has a capacitive component.

c. If shorting the load causes the minimum to shift toward the generator, the load impedance has an inductive component.

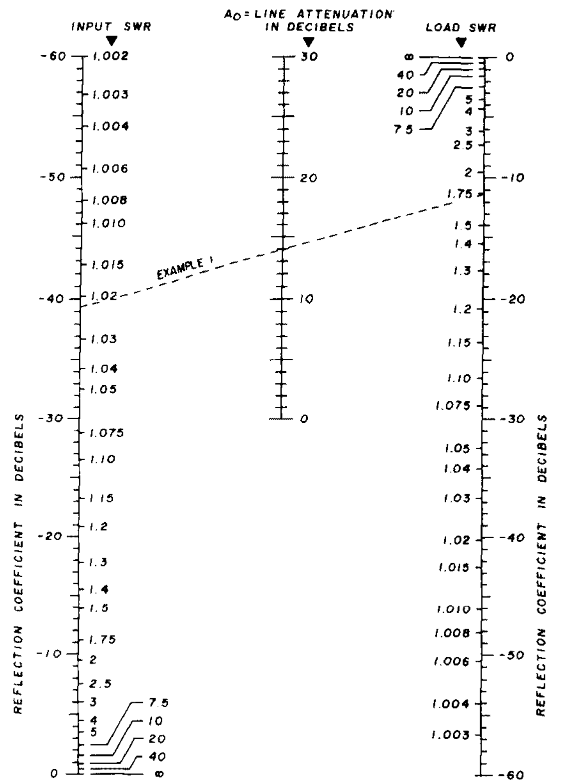



**fig. 9. Line attenuation for low input standing-wave ratios. Reprinted from reference 4 by permission of Howard W. Sams & Co., Inc.**

**d.** If shorting the load does not cause the minimum to move, the load impedance is completely resistive and has a value of $Z_o/swr$.

e. If shorting the load causes the minimum to shift exactly one-quarter wavelength, the load impedance is completely resistive and has a value of $Z_o \times swr$.

f. When the load is shorted, the minimum will always be a multiple of a half wavelength from the load.

These rules are summarized in fig. **8**.

## measuring transmission-line loss

Although the loss in a transmission line can be measured by means of an swr indicator and rf detector, as explained in reference 1, that technique requires that both ends of the line be accessible. This can be inconvenient, especially if the transmission line is connected to an antenna, and you want to find out whether the coax you put up five years ago is still good. Of course you could measure the power between the transmitter and the line, and then measure it between the line and the antenna, but this too has disadvantages (proper impedance terminations, carrying equipment to the roof or up the tower, a second person to energize the transmitter, and so on).

Let's consider how the loss can be measured with a slotted line. We know that the swr, measured at any point along a lossless transmission line, will be uniform and will depend *only* on the load impedance presented to the line. We also know that if the transmission line is lossy, the swr at the input end will appear to be better than that which is actually present at the load end. In fact, if the line is lossy enough, it will look like a pure resistance (equal to its characteristic impedance) at the input end, regardless of the terminating impedance.

Knowing these facts, it stands to reason that if the load impedance or swr is known, and the input swr can be measured, we should be able to calculate the transmission-line loss. The only problem would appear to be that of terminating the line in a known load — an antenna does not qualify — until we realize that a short is a known load with an infinite swr. Theoretically, an open circuit also presents an infinite swr, except that any connections or leads at the open end will bring the swr down.

Thus, the first step is to disconnect any existing load from the transmission line and connect a short in its place. Ideally, this should be a precision shorting termination if there is a connector at the load end of the line. If not, use a short, wide strap between the inner and outer conductors in order to minimize the inductance.

Then measure the swr at the input end of the transmission line. Obivously it should be measured at the frequency of interest because loss increases with frequency. Knowing the input swr and the load swr (∞), the line attenuation can be determined from either **fig. 9** or **fig. 10**. Those nomographs may be used for any known load; for the special case where the load swr is

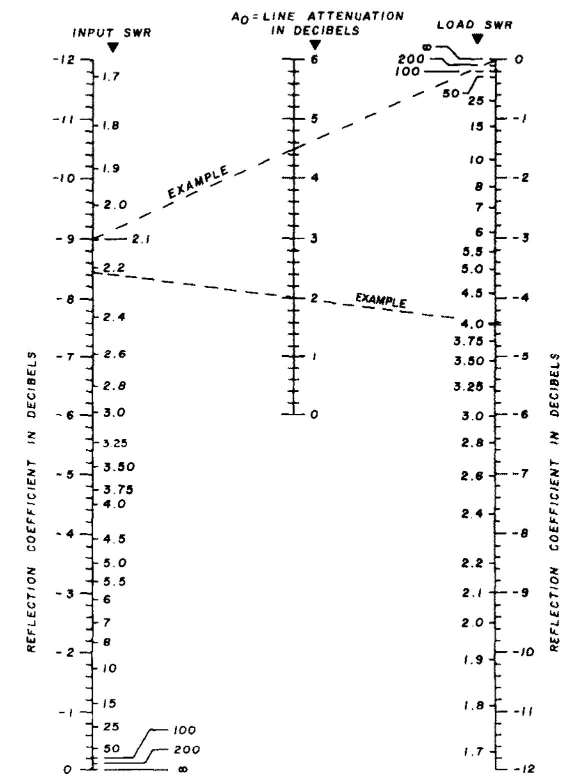
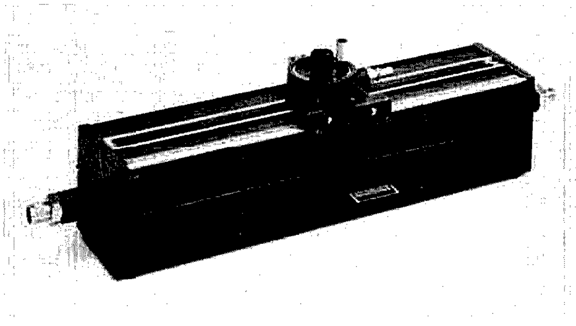



**fig. 10. Line attenuation for nigh input standing-wave ratios. Reprinted from reference 4 by permission of Howard W. Sams & Co., Inc.**



**The Hewlett-Packard Model 805C Slotted Line utilizes slab-line construction. Photo courtesy Hewlett-Packard Company.**

infinite, the following expression will also yield the attenuation, $A_o$, in dB:

$$A_o = 10 \log \frac{swr_{in} + 1}{swr_{in} - 1}$$

As an example, let's assume that you are feeding your two-meter antenna with 60 feet (18.3 meters) of RG-8A/U coaxial cable. The cable attenuation specifications, which have been plotted in fig. 11, indicate that the nominal attenuation of 100 feet (30.5 meters) should be approximately 2.3 dB at 144 MHz. However, we cannot normally use a slotted line at 144 MHz, so we must measure the loss at a higher frequency and assume that the measured loss can be translated to 144 MHz.

If we make the swr measurement at 400 MHz, the attenuation of 100 feet (30.5 meters) of RG-8A/U should be nominally 4.1 dB. Since we are concerned with the loss in only 60 feet (18.3 meters), the loss of that length should be about 2.46 dB at 400 MHz and 1.38 dB at 144 MHz.

Continuing with our example, assume that the swr measured at the input end of the coax is 2.1:1 when the load end is shorted. Going to the nomographs, it can be seen that the input swr is read more easily on fig. 10. Placing a straight-edge between the input swr of 2.1:1 and the load swr of ∞, we find that it intersects the line attenuation scale at about 4.5 dB.

Since the measured attenuation is more than 2 dB greater than what should be expected at 400 MHz, we can make a worst-case assumption that the coax has seen better days and should be replaced. If you are interested in knowing the actual loss at 144 MHz, it can be extrapolated from the curves of fig. 11 graphically, as follows.*

1. Convert the measured loss to attenuation per 100 feet (30.5 meters). Since we measured a loss of 4.5 dB for 60 feet (18.3 meters) this becomes $4.5(100/60)$, or 7.5 dB per 100 feet (30.5 meters).

*The extrapolation is based on the following approximations holding true over a *limited* frequency range: (1) the attenuation-vs-frequency characteristic is linear when plotted on log-log coordinates and (2) the attenuation-vs-frequency curve for cable having degraded characteristics varies in the same manner as that for new cable.

2. Using a scale or a pair of dividers, determine the distance between the appropriate curve and the attenuation per 100 feet (30.5 meters) calculated in step 1; this is shown as dimension $L$ in fig. 11.

3. Lay off dimension $L'$ equal to $L$, above the curve at 144 MHz; read the actual attenuation. In our case, it is approximately 4.35 per 100 feet (30.5 meters).

4. Determine the loss for the length of transmission line used. For 60 feet (18.3 meters), the loss is approximately 2.6 dB. Since the nominal loss for that length of line is only 1.38 dB, it can be seen that the line has aged sufficiently to cause an additional loss of more than 1.2 dB. Under the most favorable conditions, with a perfect match between the antenna and the coax, only 55 percent of the transmitter output power will reach the antenna, as compared to 72 percent for new line.

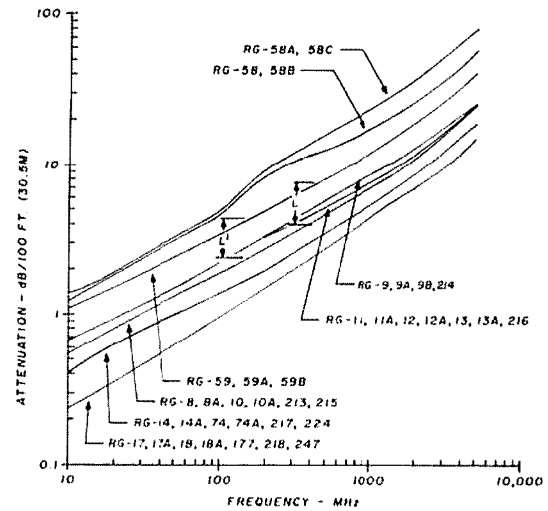



**fig. 11. Attenuation vs frequency for commonly used coaxial cables. Dimension L and L' are used in the example in the text of translating line loss to a lower frequency.**

Note that measurement of transmission-line loss is one case where a high swr is desirable, indicating low attenuation. If this seems confusing, just remember that for a lossless line, an infinite load swr will show up as an infinite swr anywhere on the line. One further comment — if the input swr actually reads infinity, it is likely that you have a transmission line which is open near the input connector, rather than a lossless line.

## references


1. Robert S. Stein, W6NBI, "Using the SWR Indicator," *ham radio*, January, 1977, page 66.
2. *Instruction Manual, Type 874-LBB Slotted Line*, GenRad, West Concord, Massachusetts.
3. *Operating and Service Manual, 805C/D Slotted Line*, Hewlett-Packard Company, Palo Alto, California.
4. *Reference Data for Radio Engineers*, 6th Edition, Howard W. Sams & Co., Inc., Indianapolis, Indiana, Chapter 24, figures 5 and 6.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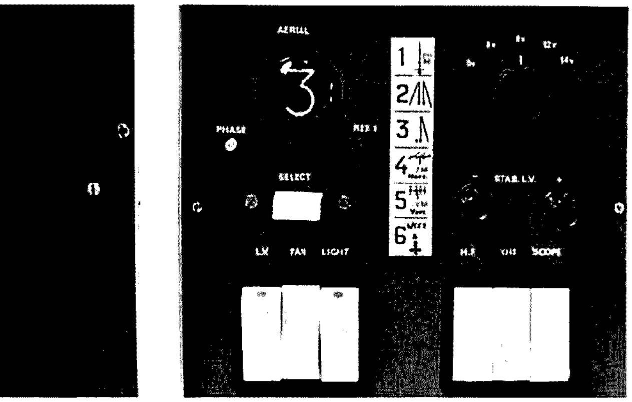


# remote switching

## multiband antennas

### Simple remote switching of amateur antennas using surplus components

This article describes a system for remote switching of multiband amateur antennas covering 2 through 10 meters. The antennas include a ten-element 2-meter beam, an inverted-V for 10-80 meters, and a ½-wave vertical for 20 meters.

The distance between my station and the antenna mast is 130 feet (40m). Adding this distance to the height of my antennas makes for a rather long transmission line. While the description of my switching setup centers around an open-wire line, coax cable could also be used, as explained later.

### system description

A single open-wire line with ½-inch (12.5mm) spacing was constructed using hard-drawn wire with polyethylene spacers located at 9-inch (230mm) intervals along the line. Near the base of the antenna mast a cast metal box measuring 6 inches (153mm) square by 3 inches (77mm) deep houses the switching circuits between the common transmission line and the antennas (fig. 1).

Switching is accomplished by a surplus Ledex rotary solenoid, which is connected to ceramic switch wafers that provide a 2-pole, 11-position arrangement. The circuit is wired to provide two 6-position repeat programs since the Ledex solenoid steps only one way (fig. 2).

A slave Ledex solenoid is used in the station to minimize the number of control wires to the switch box and to obtain a positive remote readout of switch position. To ensure synchronization between master and slave, an "indicate at position 1" line is brought into the station in addition to the two lines for energizing the Ledex coils, making three wires in all. If you wish to run an additional five wires to the station, the slave Ledex may be eliminated.

The long length of control wire to actuate the Ledex solenoid introduces a voltage drop, so power to the Ledex coil is taken from a capacitor that is charged through a resistor then discharged through the Ledex coil by a pushbutton switch. Actuating the pushbutton switch about once per second allows the capacitor to recharge; if you need more energy to work the Ledex, either voltage or capacitance must be increased. This

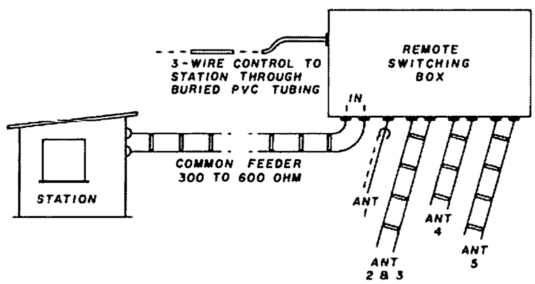


fig. 1. Arrangement of common transmission line and remote switching terminal between amateur station and antennas.

pulse method ensures that any Ledex coil can be used from 6 through 48 volts.

### display

To indicate which antenna is in use I use a nixie tube whose numerals are selected by the slave Ledex. A legend is affixed to the panel to show the antenna for each number selected.

The two Ledex solenoids can get out of phase because of an incomplete closure of the select button.

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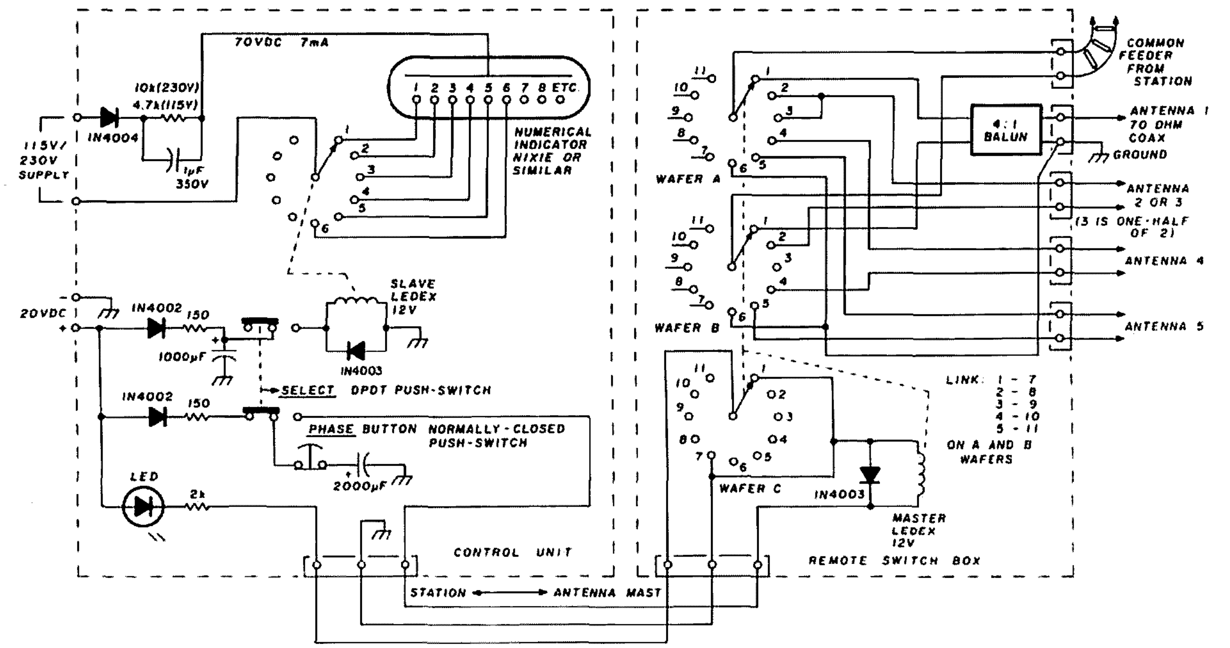


**fig. 2. Schematic of the remote antenna switching system. Master and slave Ledex solenoids are used between amateur station and antenna-mast base to minimize number of control wires. NIXIE feature shows selected antenna.**

To correct this, one control wire from the master Ledex illuminates an LED (ref. 1 in fig. 2) each time antenna 1 is selected. If any discrepancy exists, pressing the phase button disconnects the master Ledex while the slave is pulsed to read position 1 also.

## construction

The remote-control box is a surplus item. It has a terminal block on one side with feedthrough tabs. The Ledex solenoid is mounted on an aluminum bracket (see photo). Wiring is with polyethylene hookup wire. A 4:1 balun converts the open-wire line impedance (about 300 ohms) to about 70 ohms to feed the half-wave vertical antenna for 20 meters.

Wiring isn't critical, but the remote box must be weather tight. I sealed all holes and box edges with silicon rubber sealant – the type used around bathtubs and sinks, which is obtainable from most hardware stores. A small amount of silica gel was included in the remote box before sealing.

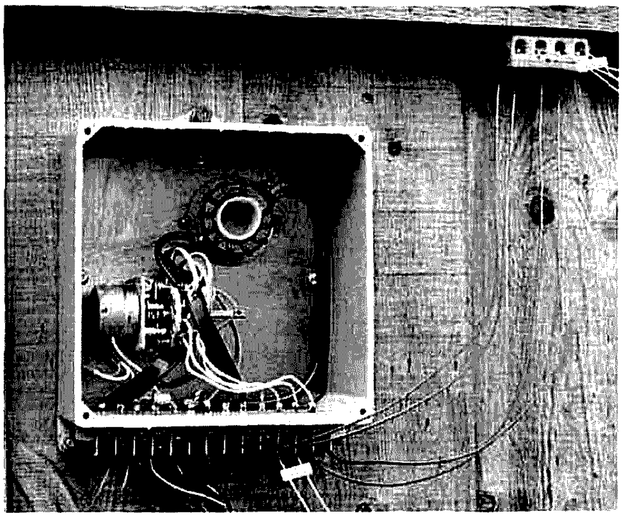

**fig. 3. Remote switch box with cover removed. Ledex stepping solenoid is shown at lower left, balun transformer at upper right. Main terminal block is shown at bottom left.**

## performance

The system described here has been in constant use for more than 7 years. Weather in England is not noted for its low humidity, but I've had no corrosion problems within the switchbox. I use the open-wire line as a tuned feeder for 10-80 meters and as a flat line for 2 and 20 meters, so the switch wafers are large (1½ inch, or 38mm). If a high-power linear amplifier is used with a tuned transmission line, contact spacing on the switch could be increased by removing alternate contacts. Using a Yaesu FT101 transmitter and tuned lines, I've experienced no problems with this system.

While open-wire line has been emphasized, coaxial cable can be used if resonant antennas are used. The low impedance and low standing-wave ratio of such antennas could be handled easily by the ceramic-switch sections. If long transmission lines are used, a good economic argument exists for the remote switching system described here.

## acknowledgement

My thanks to Bob Weston for the photos that illustrate this article.

**ham radio**

# the gin pole:

## a simple lever for raising masts

### An easy method for erecting antenna masts using readily available materials

The gin pole is a great help in raising masts. You'll find little information in the amateur radio handbooks on the use of gin poles, so in this article I've included techniques on the proper use of this simple lever to get your antenna mast off the ground and into a position where it will do some good. Information is presented on forces you can expect when using the gin pole, on rigging accessories, and on the correct way to proceed when working with the materials involved.

### the mast

A typical example is as follows. The mast butt is to be placed on the ground (or perhaps buried into the ground). If the mast is heavy, certain procedures are in order. Willing helpers can raise the mast to, say 10 feet (3m) or so above ground, then a support, such as a stepladder, can be placed under the mast. The situation is shown in fig. 1. A line secured to the point where the mast is supported by the ladder and extended in the direction you want the mast to rise can be used to hoist the mast into place, but this situation also causes some problems.

Assume the mast base is to be located at point X in fig. 1; the ladder support touches the mast at a point 25 feet (7.6m) from its base, the ladder is 10 feet (3m) high, and 100 pounds (45 kg) of weight rests on the ladder. Also assume the line is 200 feet (50m) long and attached to a tractor (or a team of helpers). A force of about 256 pounds (1139 newtons) should start the mast rising. Considering the coefficient of friction between tractor (or willing helpers) and ground surface, considerable more than this force would be required, since the force would have to be exerted nearly horizontally.

At the same time, a force of about 279 pounds (1241 newtons) acts to compress the mast between point X and the point where the line is secured. Unless the mast is blocked against movement toward the pulling force, the mast will probably be pulled off the support. To keep the mast at rest until the lift starts and to keep it from swaying later, it's best to have guy wires already attached, with helpers holding the wires to keep the mast steady.

If the mast has been properly blocked, but the mast is flexible, you may suffer the agonies shown in fig. 2 depending on where the line is secured to the mast. This effect is called buckling and the unhappy result is that the mast will suffer a permanent kink, or you'll hear a loud snap. Buckling is the most common cause of failure of long, thin masts. The answer: use a gin pole.

### gin pole number 1

There seem to be as many kinds of gin poles as people who have heard the term, and there seem to be even more ways to use them. The *ARRL Antenna Book*[1] isn't

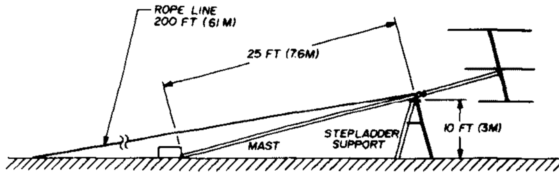


**fig. 1. A typical problem — a mast is to be erected at point X. The mast is temporarily propped with a stepladder or other support. Rope is secured to the mast as shown and extended in the direction to which the mast is to rise. Forces involved can be tremendous and the mast may buckle.**

particularly helpful, and I can name at least 20 dictionaries, encyclopedias and technical handbooks (including the *Bluejacket's Manual*) that never mention the word. One reference[2] on gin poles and ropes I've found may be available only in trade school and public libraries.

All this variety means that perhaps we'd better talk

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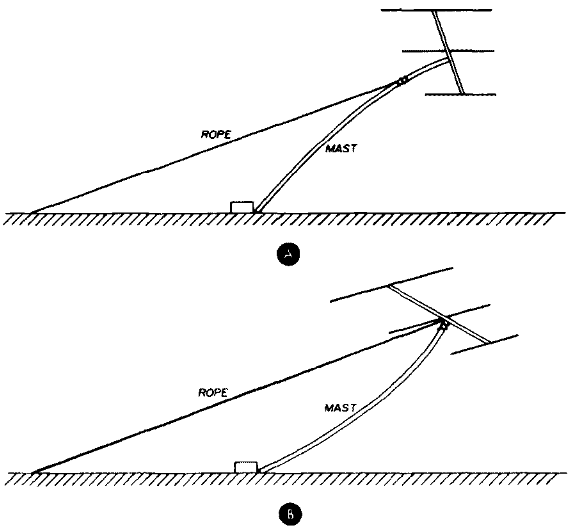


fig. 2. The buckling problem. In sketch (A) a single rope is secured above the mast balance point, which allows the mast to sag, as does a single rope secured to the top of the mast (B). In either case the mast may buckle.

about a few of the more successful kinds of gin poles so you can adapt the material you have to the job when it's time for the antenna-raising party.

Rossnagel[2] credits the American Standards Association with defining a gin pole as shown in fig. 3. (In Rossnagel's text, both weight to be lifted and pulling forces were directly below the pulley, or block). If we take the previous example but use a 20-foot-long (6.1m) gin pole located at the mast butt (point X), the required vertical force on the line would be 250 pounds (1112 newtons). The backstay would have to support 324 pounds (146 kg) and the gin pole 579 pounds (251 kg), assuming a 45-degree angle for the backstay. As the mast rises more than 53 degrees from horizontal, other people pulling on the mast guys (in the direction of the backstay) would have more effect than people pulling the gin pole line; however, the possibility of buckling still exists. Buckling would be less if one line went to our previous intermediate attachment point on the mast and another to the top of the mast. Pulling two lines at different speeds through the same block at the same time is a little like trying to pass someone when you're both going down the same playground slide. However, the two lines could be run through different blocks at the same point. (Another solution would be to run the two lines over a U-shaped bracket mounted on top of the gin pole; but the friction – and therefore needed pull – would be increased tremendously and the possibility still exists of the lines interfering with each other.

Before discussing other varieties of gin poles, something must be said about ropes and working with ropes. Rope is convenient, but the character of rope and the forces involved are dangerous to careless people.

A new Manila rope of ½ inch (12.5mm) diameter has a breaking strength of over 2600 pounds (1179kg). However, the safe working load for such rope[2] is only 265 pounds (116 kg). Such a rope would be marginal for the direct pull in the example of fig. 1 and the mast line of fig. 3. Its use as a backstay in fig. 3 would result in an overload. Doubling the rope is acceptable if both halves can be made to take the same stress. A rope should be used at about 10 per cent of its breaking strength (table 1).

Ropes that are old, rotted, kinked, wet, or frozen should be distrusted. (A wet rope or a wet splice is strong, but a wet rope kinks.) The old rules of the sailor apply: never step across a rope, never step inside a rope loop, and never wrap a rope around your hand or arm

table 1. Safe working loads for Manila rope (from reference 2). Data is based on no. 1 Manila rope, 3 strands, with a safety factor of 10.

| diameter | | working load | |
|---|---|---|---|
| inches | (mm) | pounds | (kg) |
| 0.375 | ( 9.5) | 135 | ( 61) |
| 0.5 | (12.5) | 265 | (120) |
| 0.625 | (16.0) | 440 | (200) |
| 0.75 | (19.0) | 540 | (245) |
| 0.875 | (22.0) | 770 | (349) |
| 1.0 | (25.5) | 900 | (408) |

unless you want the rope to pull you. Always use heavy gloves when working with rope lines; old ropes; particularly used ones, can produce nasty burns.

## gin pole number 2

By now it should be apparent that the gin pole is shorter than the mast. The gin pole should be 1/3 to 1/2 the height of the mast. There is an advantage (but not an

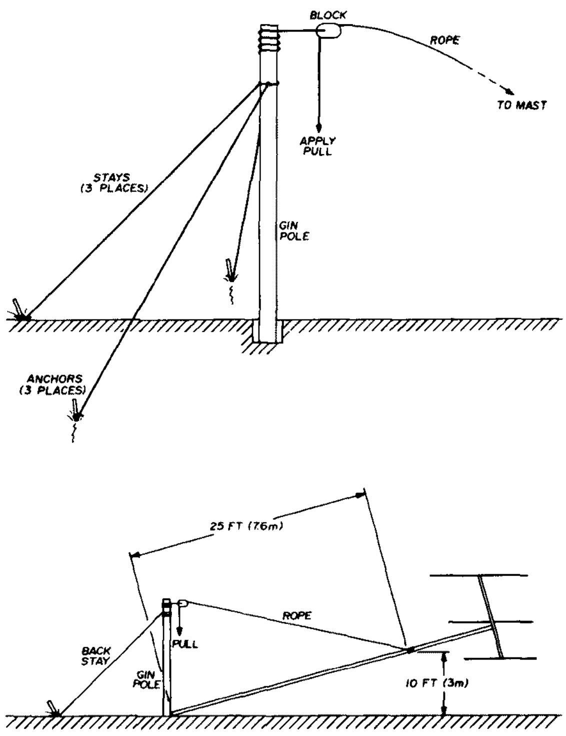


fig. 3. Gin pole no. 1: a fully guyed gin pole whose butt is located at the butt of the mast to be raised.

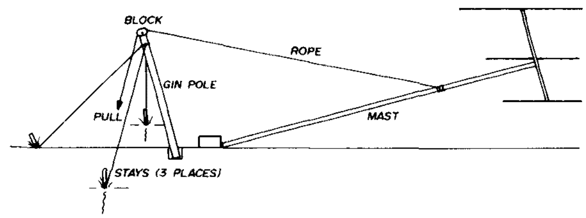


**fig. 4. Gin pole no. 2: a fully guyed gin pole located some distance from the mast.**

overriding one) in having the gin pole as high as the balance point on the mast.

In the first discussion it was assumed that the mast butt rested on the ground when the ladder support was used. If the support had been too close to the mast butt, the butt would have swung into the air, and the top of the mast would have crashed to earth. It's important that the gin-pole rope be fastened to a free mast *above* the balance point. This is what I meant when I said the butt end of the mast was supposed to be against the ground.

If the mast butt end is hinged to a heavy weight (such

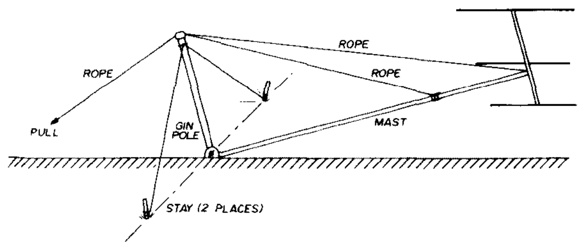


**fig. 5. Gin pole no. 3: a swinging gin pole with mast and pull ropes secured at the gin-pole top. Gin pole swings vertically. Side sway is prevented with guys or stays.**

as a foundation for the mast), the gin-pole rope line may be attached to (or even slightly below) the balance point, although I wouldn't consider that a very good idea. If the mast is of uniform cross section and the mast carries no extra load on top, the balance point is in the center. **Table 2** shows how the balance point moves on the mast as $n$ changes. ($n$ is the ratio of mast to top load, such as an antenna and rotator.) Thus if the mast weighs 4 times as much as the antenna and rotator, the balance point would be 60 per cent of the mast height. If the mast butt is larger than the top (and of the same general type of construction), the balance point would tend to be lower.

In the discussion of gin pole no. 1, pulling the rope is less effective when the mast rises to more than a certain angle. The reason is that the gin-pole rope tends to pull the mast down toward the gin pole rather than up into the air. This effect will be reduced if the gin-pole top is tilted somewhat away from the mast and the gin-pole butt likewise is moved back from the mast butt (**fig. 4**). The change of gin-pole position makes the first part of the mast raising harder but the last part easier. It does not cure the possibility of buckling, however.

While it's apparent that it's desirable to tie gin-pole ropes both at the top and slightly above the mast balance point, it may not be obvious what would be lifted by a single rope at each point in turn. On a perfectly stiff and uniform mast, a rope at the top would

**table 2. Ratio of mast weight to top-load weight as a function of balance-point location of mast height.**

| mast weight / top-load weight ($n$) | balance point location (% of mast height) |
|---|---|
| 0.5 | 83.4 |
| 1 | 75.0 |
| 2 | 66.7 |
| 3 | 62.5 |
| 4 | 60.0 |
| 5 | 58.3 |
| 6 | 57.1 |
| 7 | 56.3 |
| 8 | 55.6 |
| 9 | 55.0 |
| 10 | 54.5 |

have to lift the top load and half of the mast weight. A rope at the balance point would lift the total weight of top load and mast. Thus the rope at the top may be considered to make up for the flexibility of the mast.

## gin pole number 3

The swinging gin pole (**fig. 5**) has its bottom end pinned very close to the mast butt so that the top can swing downward as the rope is pulled. If the pull rope and two mast ropes are tied to the top of the gin pole, the mast will rise as the top of the gin pole is pulled down. The pulls at the top and slightly above the mast balance point will remain in good proportion so long as the mast butt is blocked so that the mast butt can move no further in the direction of the gin pole. (The ideal condition is when the mast butt and the gin-pole butt rotate around the same pin.) If the angle of the gin pole to the mast is 90 degrees (or a little less), the mast becomes vertical as the gin pole approaches the horizontal.

Note that the gin pole needs side strays anchored in line with the pin. This helps the gin pole to remain in a vertical plane. If the gin pole is narrow in the direction of the plane of movement, several sets of side stays along its height may help the gin pole bear much heavier loads.

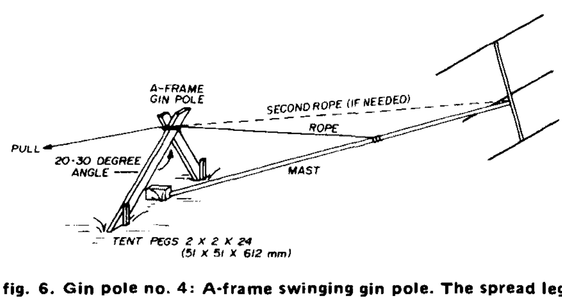


**fig. 6. Gin pole no. 4: A-frame swinging gin pole. The spread legs provide help against side sway and allow mast and gin-pole butts to line up. Tent pegs are used to butt mast and gin-pole assembly.**

Remember this as in the other procedures: helpers prevent *mast* sidesway by controlling the mast guy wires.

This technique is somewhat idealized as it isn't easy to use a single pin for mast and gin pole, and the forces on the pin can be fierce. (Since both butts can't be in the same place at the same time, the pin is subjected to bending as well as shear.)

## gin pole number 4

My preference (because of a short mast, heavy rotator, and heavy beam antenna) is the A-frame swinging gin pole of **fig. 6**. Here, two pieces of 2x4-inch (5x10cm) lumber are pinned together with a bolt, and the bottom ends of the pieces are spread into a 20 to 30 degree X. The rope is tied in the manner shown in **fig. 7**.

The A-frame legs are blocked with 2x2 inch (5x5cm) tent pegs. The tent pegs are driven into the ground in line with the pin on the hinged mast. The mast butt is likewise blocked. Each butt (mast and A-frame) should be in line with, and at right angles to, the plane of the gin pole and mast movement.

Using the swinging gin poles, the longer the pull rope (within reason), the easier will be the job of rasing the

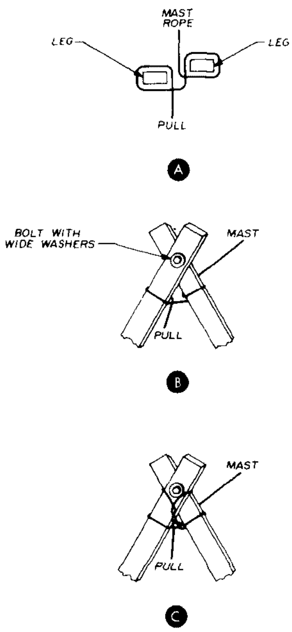


**fig. 7. Rope ties at A-frame cross. Basic method is shown in (A) and (B). Rig at (C) is preferred to minimize slipping and bind the joint.**

mast. From experience I recommend that the pull-rope length, from the gin pole to helpers, be two or three times the length of the gin pole. If a tractor (or an auto) is available to do the pulling, there's an advantage in having more rope length because of the better traction.

An auto (or any other conveniently located anchor) is more useful standing still if a long rope is available. The auto need only be moved to a location where the rope is snug. Then the pulling crew can raise the antenna by walking toward the antenna while pressing down on the rope (**fig. 8**). This procedure can bend or pull off an auto bumper, so it's best to anchor the pull rope to the auto frame.

This "walking-in" procedure is necessary at some time when raising a mast with any swinging gin pole. If *the angle between gin pole and pull rope becomes more* than 90 degrees, some of the pull force will tend to lift the gin-pole butt from its pivot. The last few degrees of

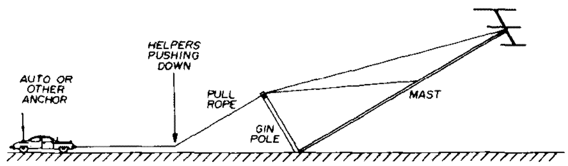


**fig. 8. The "walking-up" procedure. With pull rope tight and anchored, the pulling crew walks toward the gin pole, pushing down on the pull rope. As the mast rises, crew pulls rope during the last few feet of rise (see text).**

swing should be easy while pulling down on the rope – perhaps *too* easy! The weight of a heavy gin pole, without any pulling during the last few degrees of swing, may be enough to swing the mast to the vertical position and beyond. As the mast approaches the vertical, helpers on the mast guys away from the gin pole should keep their guys under tension so that the mast won't get out of control.

## gin pole strength

A 20-foot (6.1m) length of 2x4 inch (5x10cm) lumber isn't very strong. A finished 20-foot (6.1m) length of 4x4 inch (10x10cm) lumber will safely bear about five times the load of a finished piece of 2x4 inch (5x10cm) lumber. The larger the pole, in thickness and width, the more load it will bear. It doesn't do much good to increase one dimension of the gin pole without increasing the other, although sometimes (as with the side stays discussed for gin-pole no. 3) lateral guys will help on the narrow side.

## removing the ropes

When the mast is erect, the strain can be taken off the ropes and the gin pole can be removed. If the mast can be climbed, a helper can remove the ropes (don't forget the safety belt and its proper use). I prefer to use a heavy knob at the lift points while the mast is on the ground and use a doubled rope line from the gin pole to the lift points and looped over the knob. Thus the rope doesn't have to be tied to the pole. With two ends of the rope loose at the gin pole, one loose end is merely pulled over the knob, and the rope is down.

## references

1. *The ARRL Antenna Book*, 1st edition, 1939, page 117, or 11th edition, page 259, ARRL, Newington, Connecticut.
2. W. E. Rossnagel, *Handbook of Rigging*, 1st edition, McGraw-Hill, New York, 1950.

**ham radio**

# using a programmable calculator to design your own phased array

## How to use the hand-held programmable calculator to compute the radiation pattern of phased vertical arrays

Did you ever see those antenna pattern diagrams for an array of vertical antennas? Kraus[1] and Jasik[2] have them, and they can be found in a lot of the books the broadcast engineers have on directive arrays. Usually the books show a group of polar plots depicting what happens to the pattern when you vary the spacing and phasing of the elements. But it's not too often that they'll tell you what happens when you change the amplitudes of the element currents or lay the elements out in anything other than in a straight line. Smith's book is a noteworthy exception.[3]

You say the geometry of your back yard forces you to position your elements like a convoluted *kolbassi* sausage, and the way you want to lay your elements out doesn't look like anything in the books? How can you predict the antenna pattern? Discussed here is a little program for the HP-25 Programmable Calculator.

### the technique

A number of years ago an article appeared in the *IEEE Transactions on Broadcasting* showing how to compute broadcast array patterns when the individual towers were fed with arbitrary amplitudes and phases and were arbitrarily located in space.[4] It looked like a nice way to go if you knew FORTRAN and had an IBM 360/65 computer lying around. Now, with the advent of programmable pocket calculators, you too can crank out phased array antenna patterns — and from the comfort of your own hamshack.

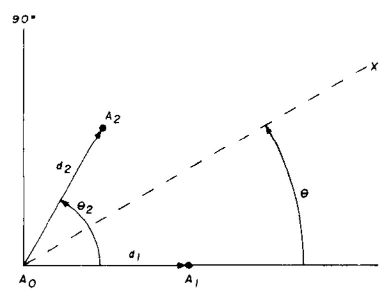


**fig. 1. Element positions for an antenna array, showing the terms used in the formula for calculating the radiation pattern. $A_0$ is the reference element.**

Suppose you lay out the elements as shown in fig. 1, choosing $A_0$ as the reference element (with current amplitude $I_0$ and phase $0°$). The magnitude and phase of the currents in antennas $A_1$ and $A_2$ will be specified as $K_1 I_0 \angle \alpha_1$ and $K_2 I_0 \angle \alpha_2$ where the $K$ and $\alpha$ values are

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## HP-25 Program Form

Title PHASED ARRAYS

Switch to PRGM mode, press [f] [PRGM], then key in the program.

| Line (Display) | Code (Display) | Key Entry | X | Y | Z | T | Comments |
|---|---|---|---|---|---|---|---|
| 00 | | | | | | | FIX 3 display |
| 01 | 2306 | STO 6 | | | | | |
| 02 | 2401 | RCL 1 | | | | | |
| 03 | 41 | − | | | | | |
| 04 | 1405 | f Cos | | | | | |
| 05 | 2401 | RCL 1 | | | | | |
| 06 | 1501 | g FRAC | | | | | |
| 07 | 2407 | RCL 7 | | | | | |
| 08 | 61 | x | | | | | |
| 09 | 61 | x | | | | | |
| 10 | 2400 | RCL 0 | | | | | |
| 11 | 41 | − | | | | | |
| 12 | 2400 | RCL 0 | | | | | |
| 13 | 1501 | g FRAC | | | | | |
| 14 | 1409 | f →R | | | | | |
| 15 | 5104 | STO+4 | | | | | |
| 16 | 21 | x⇄y | | | | | |
| 17 | 5105 | STO+5 | | | | | |
| 18 | 2406 | RCL 6 | | | | | |
| 19 | 2403 | RCL 3 | | | | | |
| 20 | 41 | − | | | | | |
| 21 | 1405 | f Cos | | | | | |
| 22 | 2403 | RCL 3 | | | | | |
| 23 | 1501 | g FRAC | | | | | |
| 24 | 2407 | RCL 7 | | | | | |
| 25 | 61 | x | | | | | |
| 26 | 61 | x | | | | | |
| 27 | 2402 | RCL 2 | | | | | |
| 28 | 41 | − | | | | | |
| 29 | 2402 | RCL 2 | | | | | |
| 30 | 1501 | g FRAC | | | | | |
| 31 | 1409 | f→ R | | | | | |
| 32 | 5104 | STO+4 | | | | | |
| 33 | 21 | x⇄y | | | | | |
| 34 | 5105 | STO+5 | | | | | |
| 35 | 74 | R/S | | | | | Stop |
| 36 | 2406 | RCL 6 | | | | | |
| 37 | 1302 | GTO 02 | | | | | |
| 38 | 2405 | RCL 5 | | | | | |
| 39 | 2404 | RCL 4 | | | | | |
| 40 | 1509 | g→ P | | | | | |
| 41 | 2407 | RCL 7 | | | | | |
| 42 | 61 | x | | | | | |
| 43 | 1409 | f→ R | | | | | |
| 44 | 01 | 1 | | | | | |
| 45 | 51 | + | | | | | |
| 46 | 1509 | g→ P | P(θ) | Phase | | | |
| 47 | 00 | 0 | 0 | P(θ) | | | |
| 48 | 2304 | STO 4 | 0 | P(θ) | | | R4 Cleared |
| 49 | 2305 | STO 5 | 0 | P(θ) | | | R5 Cleared |

| Registers | |
|---|---|
| $R_0$ | $\alpha_1 \cdot K_1$ |
| $R_1$ | $\theta_1 \cdot \beta d_1$ |
| $R_2$ | $\alpha_2 \cdot K_2$ |
| $R_3$ | $\theta_2 \cdot \beta d_2$ |
| $R_4$ | Used (cleared) |
| $R_5$ | Used (cleared) |
| $R_6$ | θ |
| $R_7$ | 1000 |

## HP-25 Program Form

Title PHASED ARRAYS Page ___ of ___

Programmer ___

| Step | Instructions | Input Data/Units | Keys | | | | Output Data/Units |
|---|---|---|---|---|---|---|---|
| 1 | Load Program | | | | | | |
| 2 | Switch to RUN | | | | | | |
| | and store element | | | | | | |
| | information | $\alpha_1 \cdot K_1$ | STO | 0 | | | |
| | | $\theta_1 \cdot \beta d_1$ | STO | 1 | | | |
| | | $\alpha_2 \cdot K_2$ | STO | 2 | | | |
| | | $\theta_2 \cdot \beta d_2$ | STO | 3 | | | |
| 3 | Store normalizing | | | | | | |
| | constant | 1000 | STO | 7 | | | |
| 4 | Enter angle at | | | | | | |
| | which P(θ) is to | | | | | | |
| | be evaluated | θ | f | PRGM | R/S | | Partial Sum |
| 5 | If N≤3 Continue | N≤ | GTO | 3 | 8 | R/S | 0 |
| | (If N>3 go to | | | | | | |
| | step 7) | | | | | | |
| 6 | Roll Down* | | R↓ | | | | \|P(θ)\| |
| 7a | If N>3 store | | | | | | |
| | element information | | | | | | |
| | on next 2 elements | $\alpha_3 \cdot K_3$ | STO | 0 | | | |
| | | $\theta_3 \cdot \beta d_3$ | STO | 1 | | | |
| | | $\alpha_4 \cdot K_4$ | STO | 2 | | | |
| | | $\theta_4 \cdot \beta d_4$ | STO | 3 | | | |
| 7b | Repeat | | R/S | | | | Partial Sum |
| 8 | Continue | | GTO | 3 | 8 | R/S | 0 |
| 9 | Roll Down* | | R↓ | | | | \|P(θ)\| |
| | | | | | | | |
| * | Now repeat Step 4 | | | | | | |
| | at the next value | | | | | | |
| | of θ | | | | | | |
| | | | | | | | |
| | | | | | | | |
| | | | | | | | |

**table 1. HP-25 program for calculating tha radiation pattern of phased arrays. Examples for its use are given in the text.**

the amplitudes and phases relative to the reference element. The antenna pattern is calculated from the phasor sum

$$P(\theta) = \left| 1 + \sum_{n=1}^{N-1} K'_n e^{j\psi_n} \right| \qquad (1)$$

where

$P(\theta)$ = the antenna pattern as a function of $\theta$

$K'_n$ = the relative amplitude of the n[th] element

$N$ = the total number of elements in the array

$\psi n = \beta d_n \cos(\theta_n - \theta) - \alpha_n$

$\alpha_n$ = the phase of the n[th] element relative to the reference element in degrees (+ for lagging phase, − for loading phase)

$\beta d_n$ = the electrical distance (in degrees) from the driven element to the n[th] element

$\theta_n$ = the spatial angle between $A_0A_1$ line and $A_n$

Because of the format which must be used with the smaller programmable calculators, eq. 1 is modified to the form

$$P(\theta) = \left| 1 + 1000 \sum_{n=1}^{N-1} K_n e^{j\psi_n} \right| \qquad (2)$$

where $K_n = \frac{K'_n}{1000}$. This may look like double talk, but it makes the program fit into the 49 steps available in the HP-25.*

*Though it may not be immediately obvious, the $K_n$ values are stored as fractional parts, thereby effectively doubling the memory capacity of the HP-25. This results in some skew in the calculated antenna patterns, but represent only a small percentage error. Editor

The program is loaded into the memory by keying in the strokes listed in the Program Form, table 1. Next you've got to put your element data into the memory registers. We're going to put in the phase and amplitude information as a decimal number of the form $\alpha_n \cdot K_n$ (with $\alpha_n$ positive). We'll also store the element position in the form $\theta_n \cdot \beta d_n$. Punch in a $\theta$ and we're ready to go. Not yet clear? Well, let's try an example.

Suppose you want to find the pattern for the array in fig. 2. Load in the program. Element 1 ($A_1$) has current $5I_o$ with phase −180°. Consequently $K_1 = 5/1000$ and we'll store the number $\alpha_1 \cdot K_1 = 180.005$ in register 0. Element 1 is along the reference axis so that $\theta_1 = 0°$, and it's spaced out $\lambda/3$ so that $\beta d_1 = (2\pi/\lambda)\ (\lambda/3) = 120°$. Therefore, store the number $\theta_1 \cdot \beta d_1 = 0.120$ in register 1 of the calculator. Next, we'll take care of element 2 ($A_2$). It's current is $2I_o \angle -90°$ so $\alpha_2 \cdot K_2 = 90.002$ is stored in register 2. Finally, $\theta_2 \cdot \beta d_2 = 45.180$ is stored in register 3. (Don't forget to put 1000 in register 7). Now you're ready to run. If you have an HP-25, preset the calculator at the program start. Let's find the value of the antenna pattern at 0° and then work our way around in 5° increments. Plug in zero, press R/S, and watch the display blink away. This program will stop at step 36 to see if you want to compute a pattern for more than three elements (more on this in the next example). Since we're looking at a 3-element array, press GTO 38 R/S.* When the program

*If you are working with three elements or less, you can save computational time by deleting steps 35, 36, and 37 from the HP-25 program, and add stack rolldown (R↓) and GTO 01 at the end of the program. This will provide direct readout of the answer without the intermediate operation as step 36. Editor

stops you're at step **49** and the display reads 0. Now rolldown the stack and read the magnitude of the field pattern at 0° $[P(0^\circ) = 5.984]$. Now plug in the next value of $\theta$, say 5°. If you repeat in 5° increments and plot on polar graph paper, you should get the pattern shown in fig. 3.

## five-element array

Next, let's calculate the pattern of an antenna with five elements. For simplicity, assume one of the arrangements you can look up in a book. Kraus provides an example on page 94 of *Antennas*: an array of five sources spaced $\lambda/2$ apart along a line. All elements are fed in phase but the relative magnitudes of the separate currents are 1, 4, 6, 4, 1, respectively (this is called a binomial array). Load the data:

$$a_1 \cdot K_1 = 0.004 \text{ (Register 0)}$$
$$\theta_1 \cdot \beta d_1 = 0.180 \text{ (Register 1)}$$
$$\alpha_2 \cdot K_2 = 0.006 \text{ (Register 2)}$$
$$\theta_2 \cdot \beta d_2 = 0.360 \text{ (Register 3)}$$

(We're using the same normalizing constant.) Plug in $\theta = 0^\circ$ and start the program. The display stops blinking with a partial sum displayed. Since $N > 3$, go to step **7a** of the procedure (table 1) and load the last two elements:

$$\alpha_3 \cdot K_3 = 0.004 \text{ (Register 0)}$$
$$\theta_3 \cdot \beta d_3 = 0.540 \text{ (Register 1)}$$
$$\alpha_4 \cdot K_4 = 0.001 \text{ (Register 2)}$$
$$\theta_4 \cdot \beta d_4 = 0.720 \text{ (Register 3)}$$

Press R/S. When the display stops blinking, you've got a

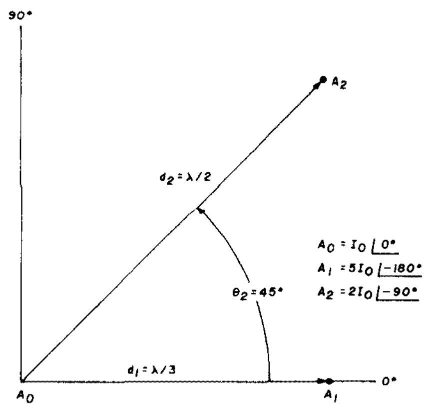


**fig. 2. Three-element antenna array used as an example in the text. Radiation pattern of this array is plotted in fig. 3.**

partial sum on all five elements. Now press GTO 38 R/S. When the lights stop, you should have 0.000 across the display. Now roll down the stack, and read off the value, $P(0^\circ) = 0.00008727 \approx 0$. Repeat the process in 5° increments and you should get the pattern of fig. 4. This

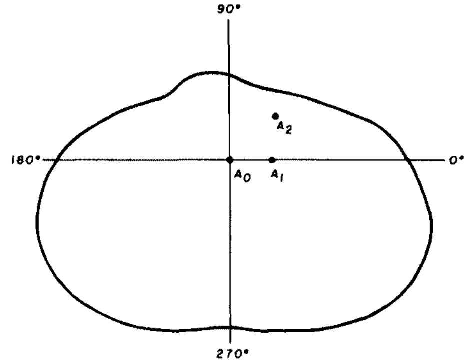


**fig. 3. Radiation pattern for a 3-element array shown in fig. 2.**

sure beats using a slide rule! If you're not convinced, take these 5 elements, lay them out at random and calculate the array pattern.

You can use this same program for an array as large as you like — just keep repeating steps **7a** and **7b** (life would be simpler if there were four more registers on the HP-25).* By the way, you can get more accuracy by letting the normalizing constant be a million. Then $R_7 = 1{,}000{,}000$; $K_n = K'_n \times 10^{-6}$; and $R_1 = \theta_1 \cdot (\beta d_1 \times 10^{-6})$.

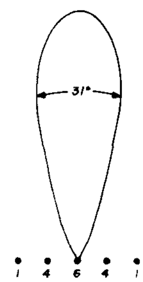


**fig. 4. Radiation pattern for a 5-element binomial array calculated with an HP-25 programmable calculator. Only the upper half of the pattern is shown.**

Remember that this gives the array pattern when the elements are fed with the *specified* amplitudes and phases. Actually, getting these on the antenna farm is no simple trick — even for two elements it can be non-trivial (see W1CF's article in *QST*)[5] and you may need some cute matching networks.

A special thanks to Bazile Pinzone of stations WCLG and WELW for several antenna sessions at the kitchen table.

* The new HP-67 and HP-97 programmable calculators have sufficient storage for up to 12 elements. Editor

## references


1. John D. Kraus, *Antennas*, McGraw-Hill, 1950, page 294.
2. Henry Jasik, *Antenna Engineering Handbook*, McGraw-Hill, 1961, see page 5.4 and Chapter 20.
3. Carl E. Smith, *Directional Antennas*, Cleveland Institute of Radio Electronics, Cleveland, Ohio, 1946.
4. E. D. Alton and L. G. Groe, "A Computer Calculation of the Far Field Radiation Pattern of an Antenna Array", *IEEE Transactions on Broadcasting*, March, 1970, pages 8-20.
5. D. W. Atchley, H. E. Stinehelfer, and J. F. White, "360° Steerable Vertical Phased Arrays", *QST*, April, 1967, page 27.


**ham radio**

# all-band bobtail curtain array

In addition to its effectiveness as a DX antenna, the bobtail curtain serves as a useful all-band antenna

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The problem of antenna height restrictions is not new to amateurs. Though, having to keep them virtually below surrounding ground can be frustrating. First the State took most of my high ground for a highway and then the FAA imposed an antenna height restriction. The tops of my new antennas varied from a point only level with surrounding terrain to about 15 feet (4.6m) above the ground level. After trying to find an effective antenna to overcome the problem, I surely felt that signals that originate in the valley, tend to stay in the valley.

My next thought was to build an antenna with enough gain to overcome the losses due to my location. Previous work with a bobtail curtain indicated it would be an effective antenna, particularly since the high current points occur at the top of the array. At the same time, the thought occurred to split the center leg into an open-wire section. The antenna could then

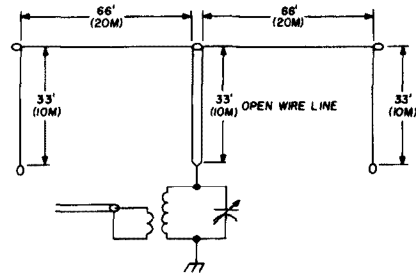



fig. 1. Layout of the all-band bobtail curtain array. The center leg has been split into an open-wire line providing a feed system for 80, 20, 15, and 10 meters. A normal bobtail curtain is one wavelength long on the top portion, with the vertical legs being one-quarter wavelength. The normal feed is through a single wire.

be fed on other bands, as a symmetrical center-fed antenna. On 80 meters, the current points would again be at the top of the array. The result is shown in **fig. 1**. I was happy to find that on 40 meters it performed exceptionally well to Australia with the temporary

25-watt rig in use. The quoted gain of 7 to 10 dB for this antenna is realistic – primarily for DX stations. For ragchewing around the United States, it is less effective. effective.

On 80 meters the tuner was wired as in fig. 2, with the antenna current distrubution shown in fig. 3. Signal reports Stateside, again, tend to be lower than reports, for instance, from the Carribean. Not enough time has been available to thoroughly investigate its properties on 80 meters.

On 160 meters the bobtail can be fed the same as on 80 and tuning will be relatively non-critical, but, the radiation angle will be higher. As a local ragchew antenna it would be quite adequate. Instead, I decided to try it working against a radial ground system. Since the resonant frequency was approximately 1.6 MHz, the antenna was "shortened" as shown in fig. 4 with about 1300 pF series capacitance. The only drawback with this arrangement is that the antenna becomes quite narrowband, requiring resetting of the series capacitor if the frequency is changed more than about 25 kHz. Radiation resistance is only 15 ohms, but this can be transformed to 60 ohms with a wideband 4:1 transformer. In my case, I used a 2-inch (5cm) diameter toroid bifilar wound with 22 turns and then connected as a 2:1 autotransformer. Using a homemade impedance bridge, the transformer looks essentially "flat". With the series capacitor resonated, the input side of the transformer measures 60 ohms. Fig. 5 is a schematic of the complete tuner, while fig. 6 shows the normal patching arrangement for 40 meters.

Although I hardly felt a 20 meter setup would be worthwhile, I arranged the tuner for 20 meters by disconnecting the 90 pF mica shunt capacitors and used the 40 meter tap on the link coil. This configuration of the bobtail begins to exhibit a pattern similar to a center-fed longwire, but also seems to produce unusually high angle lobes that may be due to interaction from the vertical ends. Since the characteristics may be partly due to the site, I would hesitate to make any definite claims regarding the various lobes, but feel it is

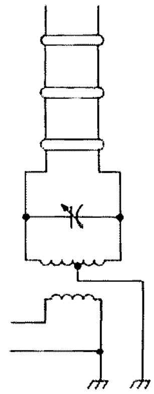


fig. 2. A single-band bobtail curtain can be fed across a parallel-tuned circuit, since the input is a high impedance. To provide a balanced feed, for bands other than 40 meters, a split secondary was used on the inductor.

a worthwhile addition – even if only as a secondary antenna. One unusual advantage at this location is the hillside to the south of the antenna effectively shields the antenna. When static from southern thunderstorms becomes severe, this antenna discriminates against the noise and provides solid contacts when the comparison antenna mounted on high ground is useless.

Since my transmitter does not cover 15 or 10 meters, it was not possible to run more than comparative listening tests. Indications are that the antenna begins to behave as a long wire on these bands, with lobes more closely aligned to the plane of the top portion. Tuning and loading were checked by using a temporary network configured the same as 80 meters, a crystal-controlled source, and a few measurements with an impedance bridge.

The antenna tuning is not critical on all bands from 80 to 10 meters. If the tuner is set for 3.6 MHz, swr does not exceed 1.5:1 over the range 3.5 to 3.8 MHz. If tuned to the center of 40 meters, the swr remains within 1.4:1 across the entire band. The higher bands follow the same pattern. With a 2-inch (5cm) graduated dial on the tuning capacitor, I can accurately preset the antenna coupler on all bands without resorting to on-the-air tuning. Thus, it is possible to have an antenna that is a solid performer on 40, but useful on all other bands with a minimum of extra effort. If 160 meter operation is not intended, no extensive ground system is necessary. A 10-foot (3m) length of pipe can be driven into the ground to act as a ground as well as to support the network. In my case, a stainless hose clamp secures the 160 meter radials to the pipe.

The complete tuning unit is installed inside of the paint pail. A tight fit prevented damage during a recent flood. The 160-meter radials are secured to the vertical support pipe by a hose clamp.

The network cover is a discarded plastic paint pail. The plywood disc was carefully fitted into the paint pail forming a skirt to guard the patch board from rain. This care in fitting paid off when an unexpected flood covered the tuner. As soon as the river was low enough, I examined the tuner and found very little

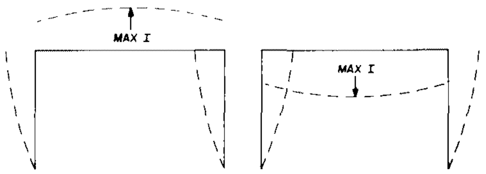



**fig. 3. Current distribution when the modified bobtail curtain is used on 80 meters.**

leakage, – only a soaked coax line. The patch board is made from a piece of plexiglass obtained as scrap. Since the input end of the feed line is a high-voltage point, this precaution is advisable. For the same reason, it is wise to "child-proof" the feed system and also the ends of the antenna. The aluminum tab

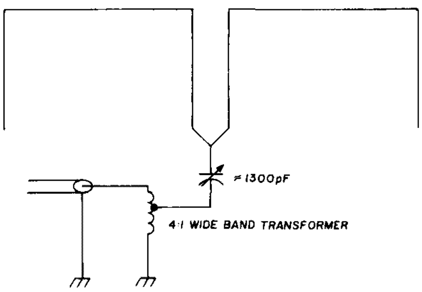



**fig. 4. Method of feeding the bobtail curtain on 160 meters. Instead of using the balanced feeders, the antenna can be fed against a system of ground radials.**

extending down from the lower edge of the plywood disc and a similar one at the rear are drilled and tapped to receive wing bolts that secure the paint pail to the disc.

As can be seen, most of the components were from a military surplus TU-5 or TU-6 tuning unit. Plate spacing is adequate for about 200 watts. As a rough guide to selection of capacitors, assuming 5000-ohm feed impedance, the following may be helpful:

| watts output | volts p-p | plate spacing |
|---|---|---|
| 50 | 1500 | 0.04 inches (1mm) |
| 100 | 2000 | 0.05 inches (1.3mm) |
| 250 | 3000 | 0.07 inches (1.8mm) |
| 500 | 4500 | 0.125 inches (3.8mm) |
| 1000 | 6500 | 0.225 inches (6mm) |

The following hints from an inveterate antenna tinker may help reduce the costs and make construction easier. Before you spend money on expensive wire, visit your nearest farm supply. Copperweld electric fence wire is cheap, light, and very strong. If you purists are turning up your noses, consider this: rf penetrates to a depth commonly given by:

$$depth = \frac{2.63 \times 10^{-3}}{\sqrt{F(MHz)}} \text{ inches or } \frac{6.68 \times 10^{-3}}{\sqrt{F(MHz)}} \text{ cm}$$

Hence, at 2 MHz, penetration is only 1.8 mils (.05mm) and becomes progressively less as frequency increases. We are then, no longer concerned with the usual cross sectional area, but more nearly with the circumference of the wire.

| wire size | circumference | radius |
|---|---|---|
| 18 AWG (1mm) | 0.126 inches (3.2mm) | 0.020 inches (0.5mm) |
| 12 AWG (2.1mm) | 0.253 inches (6.4mm) | 0.040 inches (1mm) |
| 10 AWG (2.6mm) | 0.320 inches (8mm) | 0.051 inches (1.3mm) |

The copper coating extends to a depth of typically 10 per cent of the radius. Using this fact, no. 18 AWG (1mm) would be sufficient down to 2 MHz. Thus, two strands of no. 18 AWG (1mm) are equivalent to no. 12 AWG (2.1mm) and three strands become no. 10 AWG (2.6mm).

In its basic form Copperweld is very springy, snags and kinks easily, and is difficult to twist into strands unless you use a little common sense. First, have your tools ready at the place where they will be used. A pair of trees, posts, or other supports, located the proper distance apart, will hold the wire during preparation for stranding. Be aware of the fact that when these wires are twisted you will find the group becomes several feet shorter. Fasten the end of the wire to one tree or support. Use a dowel or rod through the spool to prevent further twisting of the wire as you pay off the wire. Walk with the spool to the other tree or post, around it, and back to the start. Keep tension on the spool at all times. Your electric drill should be ready to go, with a husky hook already

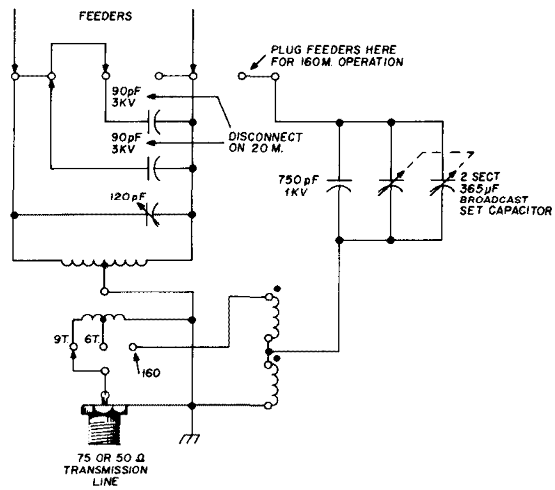



**fig. 5. Schematic diagram of the complete tuner that will permit a 40-meter bobtail curtain to be used on all bands. The feedline is plugged into the appropriate jacks for each band. The coil is 2 inches (5cm) in diameter with 17 turns (center tapped) wound to be 3 inches (7.6cm) long. The link for the input is 1-3/8 inches (3.5cm) in diameter centered inside the secondary. It should be 3/4 inch (2cm) long and wound with no. 16 AWG (1.3mm) wire. The capacitors across the coils must have adequate plate spacing to prevent arcing.**

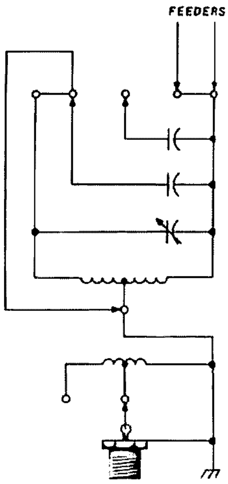
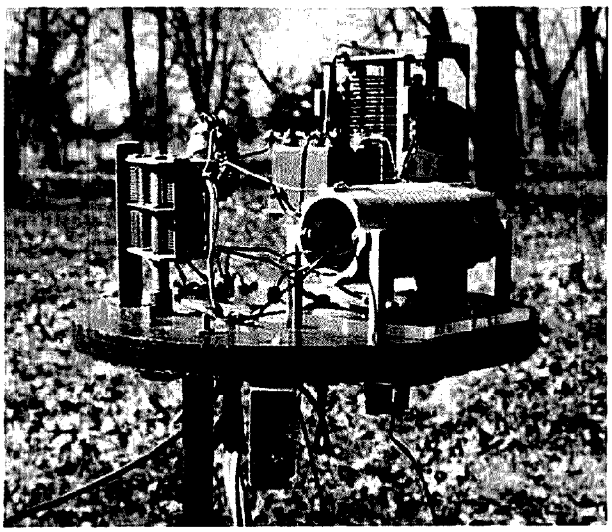



fig. 6. Patching the tuner for 40-meter operation.

in the chuck, and your cutters handy. Grasp all the strands securely and cut them free from the post. Secure the strands to the hook in the chuck. Firmly hold the drill, keeping tension on the strands to prevent them from tangling in the grass as the drill twists them. After they are twisted, and the drill shut off, you will notice the chuck rotating backwards. This is fair warning that you have made a dandy spring. If let go now, you and your lawn will be consumed by a whirling dervish, resulting in a very unhappy mess. So grasp the hook with one hand and release the chuck with the other. The added leverage from the hook will allow you to let the excess twist unwind in a controlled manner. At all times keep that wire under tension! When the wire is unwound you will have the pleasant surprise of seeing a docile, nail-straight length



Most of the components for the tuner were obtained from surplus TU tuning units. A dial on the bottom permits setting the capacitors for the different bands. The metal bracket on the bottom is used to hold the pail securely.

of wire, with no tendency to spiral or become unmanageable.

With two lengths of wire completed, you can prefab the entire antenna as shown in fig. 7. Before raising the antenna, give it a heavy coating of an acrylic spray. The best way is to make a spray guard out of tin metal stock as shown in fig. 8. Without a guard you get a light, uneven coating, and you need several cans of clear spray to do the job. The spray guard is hooked around the wire so that the rear of the guard rides on the wire. Let the wire dry thoroughly before putting the antenna up.

If you haven't priced feeder spreaders lately, get ready for a shock! However, they can be made very easily from scrap plexiglass. If just rough sawn to size, they will accumulate dirt rapidly and may arc in wet weather. I found the easiest way is to file them smooth with a sharp file, removing all saw marks. Then play the flame of a Bernzomatic torch over them. The

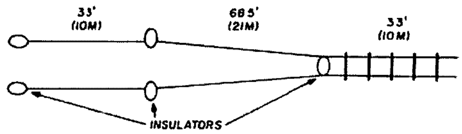



fig. 7. The entire antenna can be fabricated by installing the insulators and spreaders with the wires suspended a short distance above the ground. This will allow the feeders to hang straight.

torch will melt just the surface, leaving them glass-smooth. For powers to about 200 watts, 2 inch (5cm) feeder spacing is adequate, with spreaders spaced a foot apart. For higher power I would recommend 6 inch (15cm) spacing of feeders, and 18 inches (46cm) between spreaders. I use solid no. 22 AWG (0.6mm) wire to strap the spreaders in place and then secure these wraps to the feeders with a drop of solder. (Obviously, before spraying the wires with the acrylic spray.) By assembling the system under tension between posts or supports, the open-wire line section can be accurately made so it will hang straight.

If the array is supported between trees, you don't want the feeders bumping the ground during high winds, so your best investment is a good quality, free running pulley at one end, and an adequate counterweight. My 40 meter model, with two strands of wire, rides very steady using one half of a cement block for weight. The far end of the antenna is secured to a 55-foot (17m) tall locust tree that is about 8 inches (20cm) in diameter at the base. During high winds the point where the antenna attaches to the tree has a total swing of 8 to 10 feet (2.5-3m), yet the antenna rides very well. For inexpensive halyards on such lightweight antennas, Sears' heavy duty plastic clothesline (without the steel core) is very usable.

If you haven't worked with plastic materials, you may feel that drilling is no problem. A twist drill with a conventional grind does have a tendency to grab as it breaks through the far side. If you use too much feed pressure it will practically screw through to the far side.

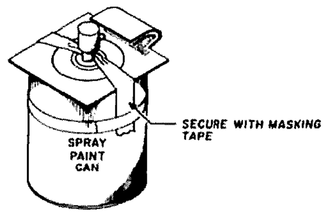



fig. 8. The guard for the spray can is fabricated from light metal stock. By using the guard, the spray will cover the wire and not dissipate in the air.

Your drill press may suddenly sport a propellor. In many cases the plastic piece will shatter or crack. An even worse hazard is present in a piece that has been squared up with a sander. The sander imparts an edge like a fine toothed saw. Let that spin on a drill press and you get an ugly slash. So break the corners and edges with a file. To prevent the drill from grabbing, modify the drill point with a brass grind. **Fig. 9** shows a conventional point, a point modified with a brass grind and the simple way to obtain the brass grind. This grind produces clean holes

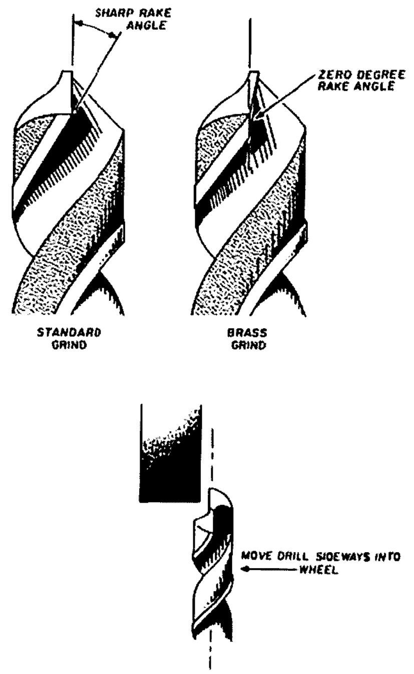
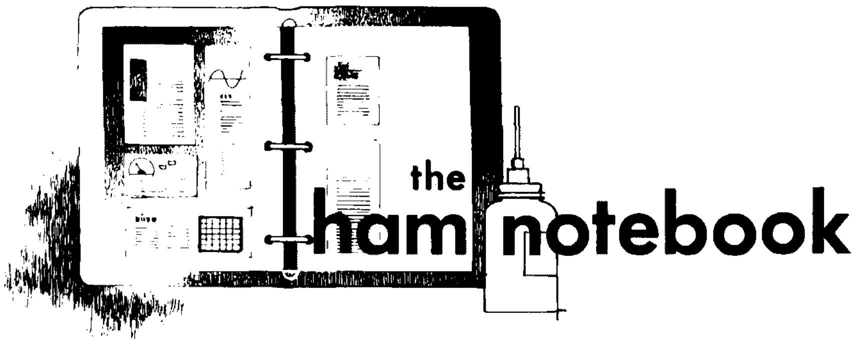



fig. 9. Normal drills can be converted to a brass grind by changing the rake angle at the tip. This alleviates problems with cracking and chipping when drilling plastics.

with far less tendency to grab unless you feed too heavily or don't retract the drill to clear chips on thicker pieces.

In conclusion, this antenna has at least gotten me back in the running! If you need a cheap, all-band antenna that can be hung from fairly low supports, is inobtrusive, and gives excellent performance for 40 meter DX, this could be one solution.

ham radio




## converting a low-band mobile antenna for two-meters

If you are a low-band mobile operator, and would like to give two meters a try without investing in a new antenna or replacing the old one, you might like to try my solution to the problem. I can convert back and forth in minutes, vswr on 2 meters is almost 1:1, and operation — either simplex or through the local repeaters — has been very satisfactory. Low-band operation is unaffected as soon as the modification is removed.

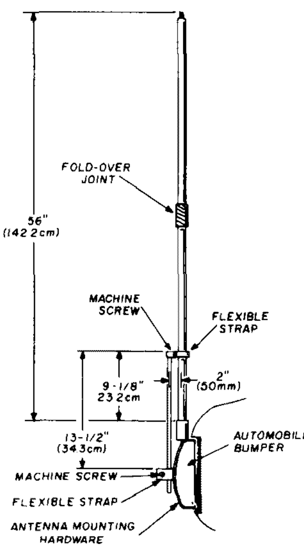



**fig. 1. Illustrates simple modification of a low-band mobile antenna to permit use on two meters. Newtronics** Hustler **is shown.**

The mast portion of my *Hustler* mobile antenna is used as the 2-meter radiator. A matching stub is made from a two-foot piece (61cm) piece of ½-inch (12.5mm) diameter aluminum tubing, a short length of flexible metal strap (plumber's tape) and two nuts, washers and bolts. The aluminum tubing serves as a matching stub held in place by the metal strap, as shown in **fig. 1**.

First, flatten the tubing for about five inches (12.5cm) at one end and bend the flattened section around the *Hustler* mast for a tight fit. Allow sufficient overlap to permit drilling the flattened portion to receive a bolt. After attachment, the tubing may be bent parallel with the mast and spaced about 2 inches (50mm) from it. Secure the bottom of the tubing to the antenna mounting hardware on the bumper by the flexible strap and a bolt, nut and washer. Adjust the distance between attachment points to about 13½ inches (34.3cm).

Dimensions for a car other than a 1973 Chrysler Newport may be a little different, but a simple adjustment of length by sliding the tubing up and down should make a good match possible.

Herb Ash, K7ARR

## vhf frequency measurement with an hf receiver and scaler

The combination of a high-frequency receiver and scaler cannot compare with a frequency counter having one Hz resolution and a scaler to measure vhf or uhf frequencies. However, if accuracy of several hundred Hz in the two-meter band is acceptable, try this.

Inject a small sample of, let's say, a 145.000 MHz signal into a 10-to-1 digital scaler. The scaled-down frequency will be exactly 14.5000 MHz. This signal, coming out of the scaler, is rich in harmonics. With the hf receiver connected to the output of the scaler, tune to the second harmonic, or 29.0000 MHz, to be compatible with the 10-meter band coverage of the receiver. In effect, this uses a divide-by-five scale-factor (10÷2=5). If the receiver dial calibration is accurate, it will measure 29.0000 MHz, showing that the vhf frequency was exactly 145.000 MHz.

The vhf frequency determination will be as accurate as the accuracy to which the hf receiver can read, times a multiplying factor of five (as in the foregoing example). My receiver dial readout error is no more than 200 Hz at midpoint between 100-kHz marker points, and about 100 Hz as it approaches the marker frequencies. This would produce vhf frequency measurements with errors of only 1000 Hz (200 x 5) and 500 Hz (100 x 5), respectively. If you are lucky enough to have the divided vhf frequency falling extremely close to a marker frequency, so that the audible beats can be counted and the marker set to a near-zero beat with WWV, an extremely good vhf frequency accuracy can be obtained — probably better than 100 Hz.

With good modern communications receivers and scalers and reasonably careful manipulation, you can obtain 500 to 1000-Hz accuracy in the two-meter band. Other combinations of scaling factors, and hf receiver frequencies, may be used to measure different portions of the vhf and uhf bands. The accuracy can be no better than the hf receiver readout times the scaling factor, but this may be accurate enough and eliminates the need for an expensive frequency counter.

Louis D. Breetz, W3LB

## boosting bargain regulators

Have you ever purchased a "bargain" 3-terminal, 5-volt, voltage regulator only to find that its output voltage was below

the manufacturer's specifications? Many of these devices provide excellent regulation, but their output voltage is marginal for TTL.

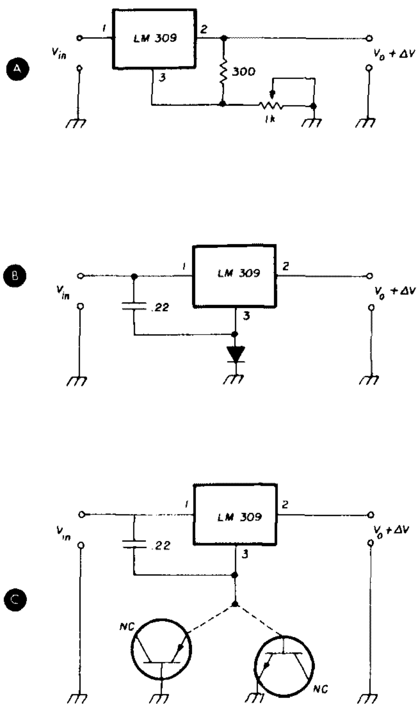



**fig. 2. Voltage-regulator circuits. (A) illustrates poor means of obtaining other than characteristic output voltage from LM309 regulator (not recommended). (B) shows a simple way to obtain a different output voltage from the LM309 while maintaining good load regulation. (C) illustrates use of a transistor to increase the output voltage of the LM309 regulator.**

The circuit of **fig. 2A** is suggested by at least one manufacturer as a way to obtain output voltages different from the device's characteristic output. Because the output voltage is dependent upon the load current, this circuit suffers diminished load regulation.

A common germanium or silicon diode as used in **fig. 2B** is a simple way to raise the output voltage while maintaining excellent load regulation. The current through the diode is less than 10 mA so a 1N270 germanium diode can be used to increase the output voltage by 0.4 volt. Any silicon diode such as a 1N4001 will raise the output voltage by 0.7 volt. The 0.22 $\mu$F capacitor is necessary to prevent oscillation when the regulator's ground terminal is raised above the circuit ground.

If you happen to have a zapped transistor on hand whose base-emitter junction is still functional, it can be used as in **fig. 2C** to boost the output voltage. Germanium or silicon transistors, npn or pnp, may be used for 0.4 or 0.6 volt increases, respectively.

Remember that the minimum input voltage to the regulator is increased by the added voltage, and the regulator case must now be electrically isolated from the chassis.

**George Shankland, WA7VVC**

## step forward in vhf circuits

**Fig. 3** illustrates a typical vhf rf stage circuit diagram, except that all of the rf circuits have been isolated from the chassis. This technique has the advantage that smaller-than-ordinary values of dc blocking capacitance can be used, helping to prevent stray out-of-band signals from reaching the mixer stage and reducing unwanted reaction between input and output circuits.

It is necessary, however, to build the rf amplifier stage and the mixer stage in separate boxes. The copper tubes which form part of the input and output coupling circuits provide a low-impedance return for rf to the coax braid, heretofore a very weak point in vhf receiver design.

All components are attached to standoff insulators mounted on each side of the central screen. The input and output coupling coils are spaced about 1/8 inch (3.0mm) from the input and output coils, respectively, to reduce their mutual capacitance to as low a value as possible while still retaining adequate coupling. Both positive and negative supply leads have high rf impedances and are bypassed to the metal box by feedthrough capacitors.

I believe that all vhf circuits should be isolated from the chassis to prevent signal coupling, and every attempt should be made to attenuate parallel currents flowing through containters (chassis, etc.) of vhf circuits. Such currents can reduce the attenuation of out-of-band signals by allowing them to bypass amplifier tuned circuits and reach the mixer stages.

Perhaps in the future parallel currents may be attenuated by painting all vhf rf chassis and containers with some form of rf-resistance paint. My copper tubes could also be covered on the outside with such paint. Although I have not tried this, it seems reasonable that higher gains, narrower bandwidths, and improved stability in all vhf rf circuits could result.

My present amplifier is similar to that shown in fig. 3, but has slightly larger values of dc-blocking capacitance and higher value rf-decoupling resistors in the supply leads. I also use a higher supply voltage of approximately 70 volts. The standoff insulators are about 1½ inches (3.8cm) long and are made from nylon. The holes through the center screen are about 3/8 inch (9.5mm) in diameter.

These techniques are also useful in reducing feedthrough of local oscillator signals to rf stages, a condition that can cause all sorts of additional problems.

**Peter W. Hazlett, G3IPV**

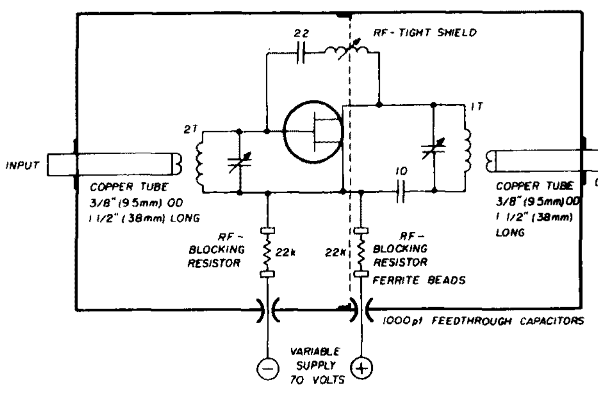



**fig. 3. Vhf rf amplifier which has been designed to eliminate problems of coupling unwanted signals into the circuit through parallel current paths.**

# short circuits

## simple computing vswr indicator

Several errors crept into the WB9CYY article on the simple computing vswr indicator, in January, 1977, *ham radio*. In fig. 2 (page 60), C9, .001 $\mu$F, should be inserted between the junction of R17 and R19. U3 is an LM301 (748), as noted in the text, while CR5 is a 1N914. The formula for calibrating the front of a junk-box meter should read:

$$swr = \frac{I_{fs} + I}{I_{fs} - I}$$

where $I_{fs}$ is the full-scale meter deflection and $I$ is the indicated meter reading. A new printed-circuit board layout is shown in fig. 1. As with all parts placement diagrams from *ham radio*, the board is shown from the component side.

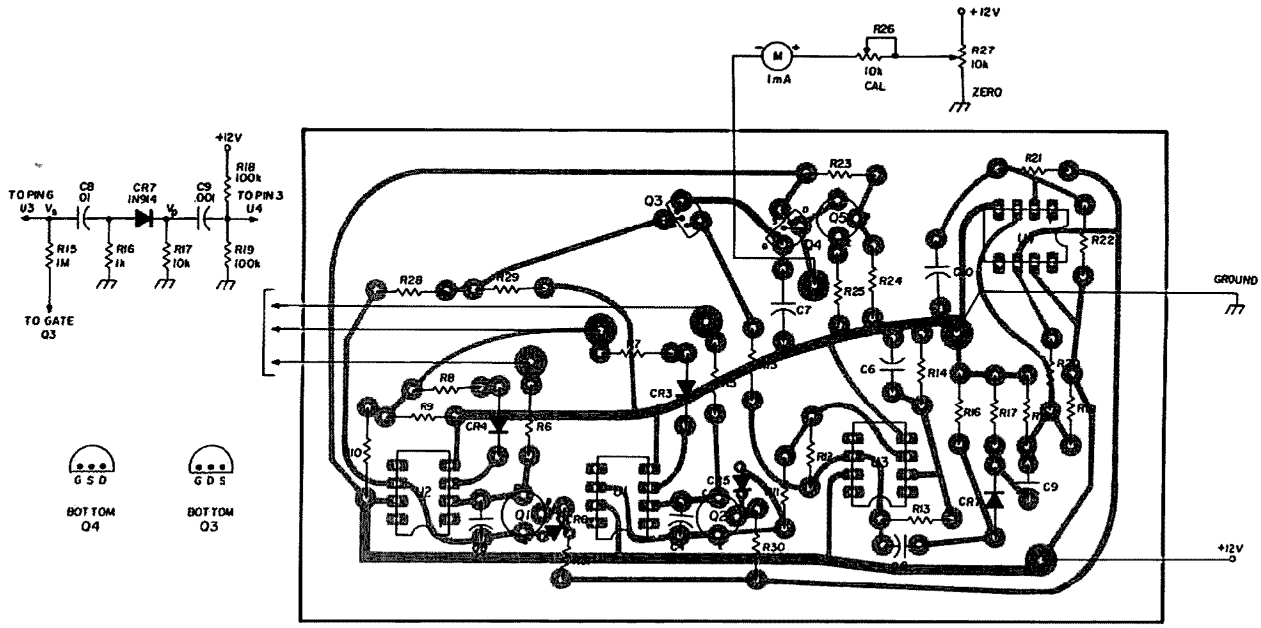



fig. 1. Parts placement and corrected board layout for the simple computing vswr indicator by WB9CYY.

table 1. Changes required to the schematic for operating frequencies other than standard mark and space.

| jumper | afsk use | Y1 (MHz) | Y2 (MHz) | R5 (ohms) | R6 (ohms) | R7 (ohms) | mark (Hz) | space (Hz) |
|---|---|---|---|---|---|---|---|---|
| 15-A | RTTY | 4.352 | 4.700 | 47k | 228 | 91k | 2125 | 2295 |
| 15-A | ASCII comp | 4.557 | 4.147 | 47k | 228 | 91k | 2225 | 2025 |
| 1-A | ASCII term | 5.202 | 4.383 | 47k | 228 | 91k | 1270 | 1070 |

$$R7 = \frac{2Q}{f_o C3} \qquad R5 = \frac{R7}{2A} \qquad R6 = \frac{R5R7}{4Q^2R5 - R7} \qquad f_o = \frac{f_{mark} + f_{space}}{2}$$

Where, $f_o > 1800$ Hz and $f_o$ = desired center frequency, Q = 10 and $f_o < 1800$ Hz, Q = 5. The actual value used for R6 should be chosen to allow variation over the desired range. Gain (A) is normally set to be = 1 (unity gain). Values in the table above were selected using these parameters.

## milliwatt portable counter

In the schematic diagram (fig. 2) of the portable milliwatt counter, *ham radio*, February, 1977, page 24, the coil shown between the output of the 40673 transistor and the input of U1 should be deleted. The network is a 330-ohm resistor in parallel with the 100-pF capacitor.

## vestigial sideband transmitter for ATV

In fig. 6 of the vestigial sideband transmitter (*ham radio*, February 1976, page 24) the input has been incorrectly shown going to pin 7 instead of pin 8. For normal operation, pin 7 is open and the video information goes to pin 8. All other connections remain the same.

## digital afsk

In the improved digital afsk (March, 1977, *ham radio*) several design changes were inadvertently overlooked by the author. Table 1 reflects these additions. The parts list should be changed as follows: C3 and C4 are .1$\mu$F mylar or polystyrene capacitors while R6 is a 500-ohm trimpot (CTS X201-R501B).

## repeater up/down mode circuit

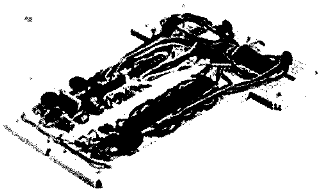
In the PC-board layout for the up-down repeater-mode control circuit, January, 1977, *ham radio*, page 41, the interconnection between pins 3 and 4 of each IC was inadvertantly left out.

## i-f amplifier design

The noise blanker for high-performance operation (fig. 2) shown in *ham radio*, March, 1977, page 11, was suggested by Siemens and developed by Michael Martin, DJ7VY.

# NEW products

## interlocking solderless breadboards from CSC



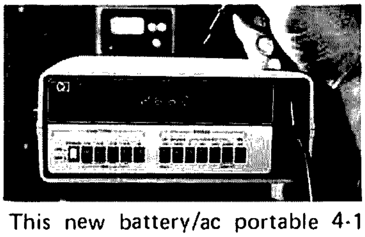
Continental Specialties Corporation, a leading manufacturer of electronic breadboarding equipment, test equipment and accessories has announced two new professional-quality solderless breadboarding sockets which combine a number of highly desirable features. Designated *Experimentor*™ *300* and *Experimentor 600*, the new one-piece sockets both provide 94 five-point terminals, plus two 40-point bus strips, for a total of 550 solderless tie-points. Experimentor 600, priced at $10.95 suggested list, has a 0.6" (15mm) center channel, making it the only socket on the market with full 4-terminal fan-out for microprocessors, clock chips, RAM's, ROM's and other larger DIP packages. Experimentor 300, priced at $9.95 suggested list, has a 0.3" (7.5mm) center channel that is perfect for smaller DIP's.

Like CSC's other popular breadboarding products, both Experimentor sockets also accept transistors, LED's, resistors, capacitors, pots — virtually all types of discrete components, as well as lengths of # 22-30 solid hookup wire for interconnection — with plug-in ease. CSC Experimentor sockets also feature a unique interlocking system that permits sockets to be snapped together, mixed or matched, vertically or horizontally, to provide optimum configurations for almost any type of circuit. And instantly disconnected or reconnected, without tools, to meet requirements.

CSC Experimentor sockets are molded of durable abrasion-resistant material, and feature CSC's non-corrosive, prestressed nickel-silver contacts for positive connection and longer life. Both sockets measure 0.325" (9.5mm) deep and 6" (15cm) long — but the Experimentor 600 measures 2.4" (6cm) wide (as opposed to 2.1" (5.3cm) for Experimentor 300), because of the wider center channel that accommodates microprocessor type DIP IC's. Both sockets also feature alphanumerically designated tie-points, aiding in circuit design and testing, as well as circuit tracing. Vinyl-backed to prevent accidental shorts, CSC Experimentor sockets can be used free-standing or conveniently screw-mounted, either with 4-40 flat-head screws from the front, or 6-32 self-tapping screws from the rear.

CSC Experimentor sockets are available now from CSC distributors and dealers, or directly from CSC's East- or West-Coast offices. For more information, contact CSC at 44 Kendall Street, Box 1942, New Haven, Connecticut 06509, or 351 California Street, San Francisco, California 94119.

## digital multimeter



This new battery/ac portable 4-1/2 digit, five-function digital multimeter from Hewlett-Packard has a unique *touch-hold* probe (available as an accessory) that lets the user "freeze" the reading on the display — a convenience when probing closely-packed circuit boards.

Called the Model 3465B, the digital multimeter has a dc voltage measurement range from 1 microvolt to 1 kilovolt with a mid-range accuracy of ±(0.02 per cent reading + 0.01 per cent of range) for one year. Ac measurement range is 10 microvolts to 500 volts with a mid-range accuracy of ±(0.15 per cent of reading + 0.05 per cent of range) over a 40 Hz to 20 kHz bandwidth.

Ac and dc current measurement range is from 10 nanoamps to two amps. Dc current accuracy for the 10 mA range is ±(0.1 per cent of reading + 0.01 per cent of range). Ac current measurements are made over a frequency band of 40 Hz to 20 kHz with a mid-band accuracy of ±(0.25 per cent of reading + 0.25 per cent of range).

Resistance range is 10 milliohms to 20 megohms with a mid-range accuracy of ±(0.02 per cent of reading + 0.01 per cent of range). Open circuit voltage on the *ohms* terminal, when set to its lowest range does not exceed 5 volts, preventing damage to most solid-state devices.

Input protection is provided to 1kV on any dc range, 500 V rms on any ac range, and 350 V peak on any resistance range. A front-panel fuse protects the instrument from overload when measuring current.

The HP 34112A touch-hold probe accessory provides greater utility by allowing the operator to focus his attention on the point of measurement in hard-to-reach circuits. The touch-hold probe, which plugs into the front panel input connectors, holds the displayed reading at the touch of a button.

A high efficiency LED display has the advantage of longer battery life, and instrument reliability is improved because of low internal temperature rise.

Input terminals are recessed to meet safety requirements, and the input terminal for current also contains a fuse. International symbols as well as voltage limits are shown on the front panel.



More Details? CHECK—OFF Page 150

The standard 3465B DMM comes with an internal ac power supply and rechargeable nickel-cadmium batteries.

U.S. price of the Hewlett-Packard Model 3465B with rechargeable nickel-cadmium batteries is $500. U.S. price of the Model 34112 Touch-Hold Probe is $40. For more information write Inquiries Manager, Hewlett-Packard Company, 1501 Page Mill Road, Palo Alto, California 94304.

## Hamtronics catalog

Hamtronics announces publication of a new catalog featuring, among other products for the vhf enthusiast, a new miniature vhf receiver preamplifier, a receiver multi-coupler allowing two receivers to be used on a single antenna, and a low-cost fm signal generator. Previous products from *Hamtronics* listed in their new catalog include vhf and uhf fm receivers and transmitters in kit form and various adapters for use with vhf and uhf equipment, such as scanner adapters, multi-channel adapters, and a full line of preamps.

The new 16-page catalog is yours in exchange for a self-addressed stamped envelope. Write Hamtronics, Inc., 182 Belmont Road, Rochester, New York 14612.

## temperature-controlled soldering iron

The new *Oryx 50-3* soldering iron has a thermostatic control built right into the handle. It keeps the temperature constant regardless of speed of soldering, or line voltage variations. Operating temperatures can be adjusted in seconds to any setting between 400° and 750° F (204° and 399° C) while the iron is running. An indicator light in the handle serves as a visual guide to the control system, as a temperature-setting aid, and as a safety feature by indicating clearly when the iron is left on. The *Oryx 50-3* comes complete with a long-life, iron-coated tip and a four-foot, 3-wire cord for 100-130 Volts ac. Other style tips and a safety stand are offered as optional accessories. For more information, write: Oryx, 4115 North 44th Street, Phoenix, Arizona 85018.

$1.50

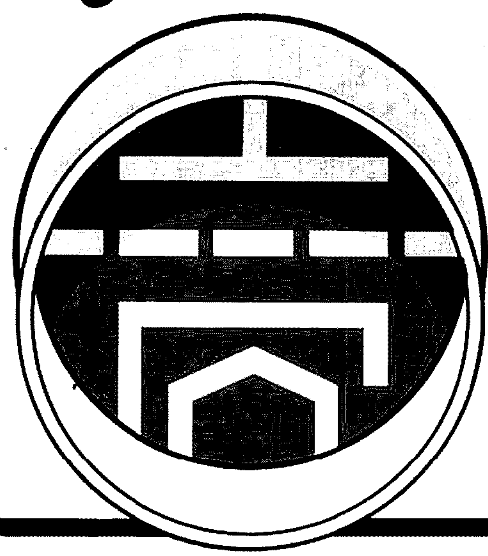
# ham radio

magazine

## JUNE 1977



stripline 432-MHz power amplifier



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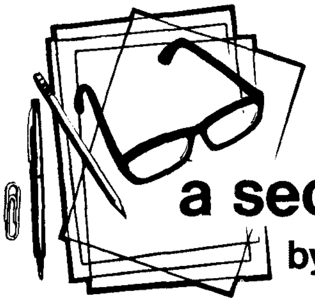


# a second look

by Jim Fisk

In early March the FCC released new rules which could have a substantial effect on the whole future of amateur radio. The new rules, spelled out in Docket 20777, became effective on April 15th, and require that the spurious emissions from *all high-frequency* transmitters, transceivers, and amplifiers must be at least 40 dB below the mean power of the fundamental, without exceeding 50 milliwatts. The rules for vhf transmitters operating between 30 MHz and 235 MHz are even more stringent: for transmitters of 25 watts or more, spurious emissions must be down a minimum of 60 dB. Note that this is not just a proposal — it is the law, and it applies to *all* amateur equipment: new, used, or presently in use!

Furthermore, as the law is now written, there is no provision for the use of external filters to reduce spurious emissions to the required level; the rules imply that the required purity of emissions must be measured at the transmitter or amplifier output connector. Recent tests by the ARRL Technical Department with a Hewlett-Packard spectrum analyzer indicate that *most* of the commercial high-frequency transmitters now on the market meet the new requirements.

In most cases, when the Commission adopts new rules, it makes an effort to minimize hardships and harmful economic effects on the licensees — usually by providing enough lead time so equipment can be brought into line with the new requirements. That was not done in this case. Most amateurs, in fact, probably weren't even aware of the new restriction until several weeks after it went into effect, and few of those who did know about it had any way of measuring the spectral purity of their transmitters!

All amateur equipment sold after April 15th must comply with the new restrictions on purity of emissions; but what do you do if you are using equipment which was manufactured before April 15, 1977? The law doesn't tell you *how* to comply, only that you must. The best way is with a spectrum analyzer, but a new one costs as much as a small house, and good used ones cost as much as a complete transceiver. You might consider the homebuilt analyzer described in this issue (and we'll have others in the future) or perhaps the members of your club can be persuaded to pool their assets to buy one. There are also commercial test labs which can make the measurements for you, but no matter which approach you take, the price isn't going to be cheap.

The ARRL has petitioned the FCC to stay the effective date of the new rule for nine months, and they have also petitioned for a reconsideration of the action, but at this point it's impossible to tell what the outcome wvllnbe. In the meantime, all amateur transmitters must comply with the new restrictions.

The new rules on spectral purity are just one of a series of restrictions on amateur equipment which have been brought about by unscrupulous CB dealers and manufacturers who are peddling amateur transceivers and illegal broadband "amateur" linears to CBers. The FCC district offices are besieged with complaints about RFI from these devices; in some areas television interference from CB is extremely severe, and there's a strong feeling in Washington that the CB situation is now so far out of hand that drastic action is necessary.

One way the FCC feels they can control this is to place an outright ban on the manufacture of linear amplifiers which operate between 24 and 35 MHz, and to require type acceptance of all commercial amateur transmitters. Neither of these actions would probably have much effect on the illegal CB operators (any more than gun control removes guns from the hands of criminals), but it will surely increase the prices you'll have to pay for your amateur equipment in the future. It will also mean that the manufacturers won't be able to offer you new circuits and technology as fast as they have in the past — each improvement they make in their equipment will require a whole new round of type acceptance. In-line production improvements will cease, and the natural evolution of modern amateur gear will grind to a halt.

In a statement that accompanied Docket 20117, the amplifier ban, FCC Chairman Richard Wiley expressed the hope that "the comments we receive will suggest other and better alternatives to the Commission's proposals." One alternative, proposed by the San Antonio Repeater Organization (SARO) would place legal responsibilities on the seller and buyer of amateur transmitting equipment. The R. L. Drake company has strongly endorsed the SARO proposal, but has worked out a modification which would remove the paperwork burden from the FCC and still provide the Commission with traceability and accountability for enforcement.

The Drake plan, which is based on presentation of a valid amateur license when purchasing transmitting equipment, is especially recommended. Under this plan your callsign would be recorded on the sales invoice along with the equipment serial number. If and when you sold the equipment you would keep a record of the amateur you sold it to. This plan would not add to the cost of amateur equipment because it's based largely on records which are already being kept as part of good business practice.

If the FCC located amateur equipment in the hands of an unlicensed operator, it would be a simple matter to trace the equipment, by serial number, from the manufacturer to the point where it crossed over from amateur to CB. The violator would be subject to fines up to $500 per day during the time which the offense occurred.

This proposal requires your support — the whole future of amateur radio demands it. Let's place the burden of the CB problem where it belongs: on the shoulders of the unscrupulous dealers and illegal operators.

**Jim Fisk, W1HR**
**editor-in-chief**

# presstop

A PROPOSAL TO LIMIT TRANSMITTER SALES to Amateur licensees was filed with the FCC by the San Antonio Repeater Organization in response to an FCC request for workable alternative solutions to the problem of non-Amateur use of Amateur linear amplifiers in the CB bands (May HR). The SARO proposal, assigned RM-2839, drew a tremendous — possibly record-breaking — number of Comments filed in its support.

R.L. Drake And ARMA were both among those filing strong supporting Comments on RM-2839. Drake's offering even described a procedure for implementing a "Proof of License" program which would require little investment of time or money by the FCC.

FCC/AMATEUR DIALOGUE was severely curtailed as a result of a court decision handed down recently in Washington. In its ruling in the case of Home Box Office vs the FCC, the U.S. Court of Appeals of the District of Columbia stated:

"Once a Notice of Proposed Rule Making has been issued...any agency official or employee who is or may reasonably be expected to be involved in the decisional process of the rule-making proceeding, should refuse to discuss matters relative to the disposal of the Rule-Making proceeding with any interested private party or any attorney or agent for any such party prior to the agency's decision."

This Prohibition Is Being interpreted to extend to 60 days after the final Report and Order on an NPRM becomes effective, to include the period in which a Petition for Reconsideration may be filed. An appeal of the decision by the FCC is considered likely but could in itself take years.

CONFUSION BETWEEN "TYPE ACCEPTANCE" and "Type Approval" seems to be contributing lots of heat but little light on much of the current discussion of Docket 21117, the FCC's proposal to require commercially-made Amateur transmitters to be Type Accepted. Type Acceptance requires only that the manufacturer submit certain performance data on his product and certify that what he markets will meet or exceed the performance he claims for it. Type Approval is a much different process, with the Commission itself performing elaborate tests following extensive testing by the maker.

As One High-Ranking FCC Spokesman said, "The FCC Type Acceptance procedure doesn't ask for any data that a manufacturer shouldn't have developed for himself, long before he was ready to put his product into production." This isn't to say that the FCC could not, or would not, require additional testing of suspect equipment in its own labs — that's what happened to CB, after FCC found that practically no CB sets met claimed specifications.

JACK ANDERSON SLAMMED AMATEUR RADIO, and the many dedicated FCC people who are also Amateurs, in his nationally-syndicated column that appeared in many papers Monday, April 4th. Thrust of the piece, which cited confusing comparisons from a confidential report prepared for U.S. Representative Elliott Levitas (D-Ga.) was that letting Commission Amateurs make CB policy was letting "the wolf guard the flock" and implied ARRL membership might be a conflict of interest for FCC personnel.

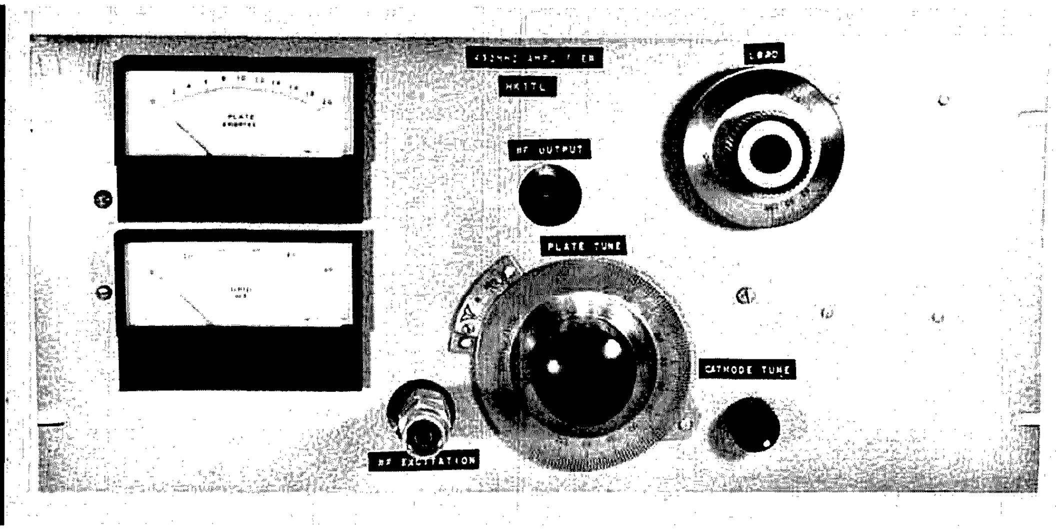
The Column Turned Out to have provided one of the best PR opportunities Amateur Radio has had in recent years. Copies of many letters to the editor, a good number of them already published, have been received by HR — and Pete O'Dell, ARRL's Public Information Officer, says the League has a pile of pro-Amateur Radio clippings about four inches high with more coming in every day. In Michigan WB8VBP even managed to turn Anderson's blast into invitations to two radio talk shows, where she was able to push the positive side of the Amateur Service very strongly.

"HOW TO RESOLVE RADIO-TV INTERFERENCE PROBLEMS" is a new 35-page booklet by the FCC due out sometime in June. Coverage includes interference to telephones and a variety of electronic equipment, with useful tips for both homeowners and service technicians. Cost is about $1.50 from the Government Printing Office.

RUSSIAN COMMERCIAL AMATEUR GEAR was to have been produced for the first time under a DOSAAF (a somewhat MARS-like organization) five-year plan begun last year. Transmitters and receivers are already supposed to be out, with transceivers, keyers and other accessories to be introduced this year.

THE U.S. AMATEUR POPULATION passed 300,000 for the first time during March, with 300,372 licensed U.S. Amateurs at the month's end. The total at the beginning of the month was 296,967, and 6,693 new licensees during March more than doubled the 3,288 who let their licenses expire during the month. A year ago we numbered only 265,528, so we increased almost 35,000 in one year!

ALIEN AMATEURS WHO become citizens are no longer eligible to hold reciprocal licenses, even if their non-U.S. licenses are still valid. The new citizen must take the U.S. Amateur examination and receive a U.S. license if he wishes to remain on the air, according to a recent FCC release.



# 432-MHz power amplifier using stripline techniques

Design and construction details for building the rf amplifier used in the EME expedition to South America in 1976

This article describes a 2-kW peak-envelope power (PEP) amplifier for 432-MHz using the Eimac 8938 triode in a stripline configuration. The design evolved from work previously done by K2RIW[1] and W3CCX. This amplifier was used during the "Pack Rat's" earth-moon-earth (EME) expedition to South America in July and August of 1976. The success of the expedition was partly due to the reliability of the equipment, which included this amplifier.

Kilowatt amplifiers for 432-MHz aren't available at your favorite hobby shop. Amateurs who like to work EME either build amplifiers from scratch or from kits. A popular solution to the kW amplifier for EME is the design described by K2RIW. His stripline technique was adapted to the amplifier design described here.

## background

Tests of a 432-MHz kW amplifier loaned to the South America expedition members before the trip indicated that power output suffered from poor plate-current efficiency. Worse, maximum output power was marginal for EME work, leaving no reserve in the event of low driver output or low supply voltage. This situation often occurs in a remote location. We decided to proceed with a new design for an amplifier for our South America operation.

## design approach

Trouble-free operation during the planned 12-hours-per-day operating schedule, using slow CW (high duty cycle), dictated a military-type approach to amplifier design. Such an approach is to use a device rated at twice

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the expected requirements to ensure reliability. An amplifier used by W3CCX uses an Eimac 8938 triode in a grounded-grid, cathode-driven configuration. This tube is rated to 500 MHz with 1500 watts plate dissipation. The tube is a coaxial-base version of the popular 8877 triode.

Advantages of the 8938 amplifier at W3CCX were trouble-free operation and easy drive requirements using only a straightforward power supply; a disadvantage of the W3CCX amplifier was that the cavity construction required metal-working facilities. Also, output-loading adjustment was difficult and time consuming. The only way to couple output from a cavity is through an inductive link. Those who have experimented with link coupling in hf gear know the frustration involved. These frustrations are even worse in vhf cavity designs because of the limited number of adjustments possible.

Recently published vhf amplifier designs have used stripline techniques with excellent results. Striplines have been used from 50 through 1296 MHz. The ease of tuning and loading the K2RIW 432-MHz stripline amplifier convinced me that this was the way to go; what remained was to adapt the technique to the 8938 coaxial-based triode.

## description

The plate circuit is a half-wave stripline with flapper capacitor tuning and loading controls. The grounded-grid triode is cathode driven using a half-wave stripline cathode circuit tuned by a movable capacitor disc. Matching to the driver transmission line is through a variable capacitor also of a movable-disc type. The amplifier enclosure consists of two aluminum sheet-metal chassis boxes similar to the K2RIW amplifier with flat-plate top, middle, and bottom plates. The amplifier is easily built using simple tools and lends itself to simple disassembly for inspection and parts replacement. The only purchased parts were a blower, some brass shim stock, and a few nuts and bolts. The remainder of the parts were either adapted from the junk box or scrounged from friends.

## design and construction

The amplifier schematic is shown in **fig. 1**. It is a straightforward grounded-grid triode which is cathode driven using zener bias. The operating voltages required are B+ and filament — period! The plate stripline is located in the upper box 12 inches long x 7 inches wide x 4 inches deep (306x179x102mm), and the cathode circuit is located in the lower box 7 inches long x 5 inches wide x 3 inches deep (179x128x77mm). A 1/8 inch thick (3mm) aluminum center plate divides the two chambers and provides mounting points for the tube socket and the stripline standoff pillars.

***Plate circuit.*** The plate circuit is a half-wavelength stripline made from a 5 x 8 inch (128x204mm), double-clad, 1/8-inch-thick (3mm), glass-epoxy PC board. The corners are rounded to minimize voltage discontinuities. The stripline is located approximately midway between the compartment baseplate and cover, producing a transmission line with a characteristic impedance, $Z_o$, of about 56 ohms. Capacitor C3 (about 0.5 pF at resonance) tunes the line. The dimensions of C3 are 3 inches (77mm) wide and 3¼ inches (83mm) long. The 1/32-inch-thick (1mm) brass shim stock capacitor overlaps the outer ½ inch (12.5mm) of the plate line.

Capacitor C4 is a beryllium copper flapper ½ inch (12.5mm) by 2 inches (51mm) long, which is soldered at one end to the center contact of the type-N output

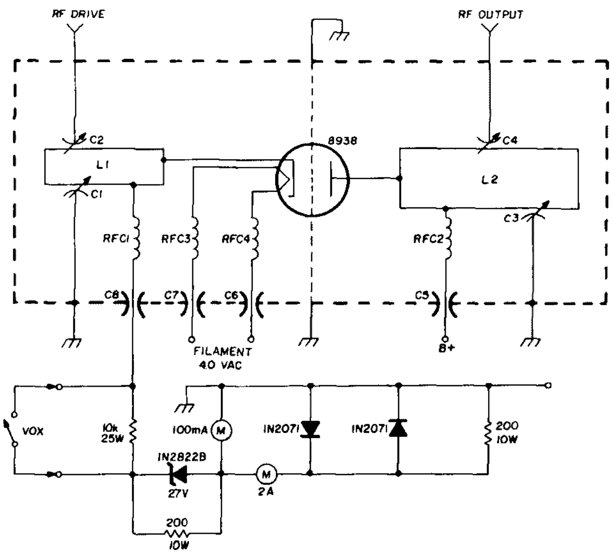



**fig. 1. 432-MHz amplifier schematic using stripline construction. Nominal 13-dB gain is provided with this design, which emphasizes reliability for operation under adverse line voltage and temperature conditions.**

| | |
|---|---|
| C1 | 1 inch (25.5mm) disc on piston tuner |
| C2 | 1 inch (25.5mm) disc mounted on transmission line center conductor |
| C3 | Brass shim stock flapper (see text) |
| C4 | Beryllium copper flapper (see text) |
| C5-C8 | Erie 1000 pF feedthrough bypass capacitors |
| L1 | 1/16-inch-thick (1.5mm) brass sheet cathode line. |
| L2 | Double-sided PC board plate line |
| RFC1-RFC4 | Rf chokes (see text) |

connector. A ¼-inch-thick (6.5mm) plexiglass block bolted to the plate compartment side wall prevents rotation of the output coupling capacitor and type-N connector center pin. Capacitor C4 overlaps the plate circuit by 1 inch (25.5mm).

Both C3 and C4 are positioned by dial-cord strings. The string for C4 passes straight down from the capacitor through a clearance hole in the base plate to a ¼-inch-diameter (6.5mm) steel rod, which is rotated by the plate tuning knob. This control is a National *Velvet Vernier* planetary drive reduction unit. The output capacitor string leads directly upward from the capacitor, passes over a standoff-mounted fairlead, and exits the compartment at the end. It then wraps around

a ¼-inch-diameter (6.5mm) steel rod, which is also driven by a National *Velvet Vernier* knob. The capacitors are retracted by the pull of the strings and return due to spring action when the string tension is relaxed. Action is smooth and positive. A string with a good dielectric must be used. Some fly fishing line was tried but melted in the strong rf field!

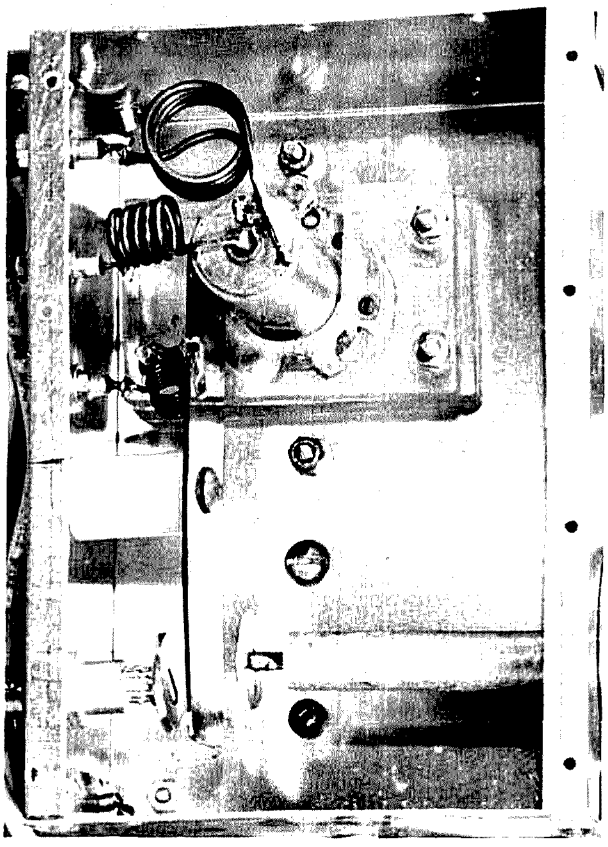
Cathode circuit. The cathode circuit is a half-wavelength stripline counterpoised against one wall of the cathode compartment; it is tuned to resonance by a disc capacitor to ground at the open end of the line. Near this same end of the line, another disc capacitor couples drive power into the cathode.

The cathode line is made of 1/16-inch-thick (1.5mm) brass sheet 4½ inches (115mm) long by 1¼ inch (32mm) wide, which is soldered at one end to the tube socket cathode terminal. A teflon pillar, 3/4 inch (19mm) diameter by 1¼ inch (32mm) long, supports the cathode line at about its physical midpoint.

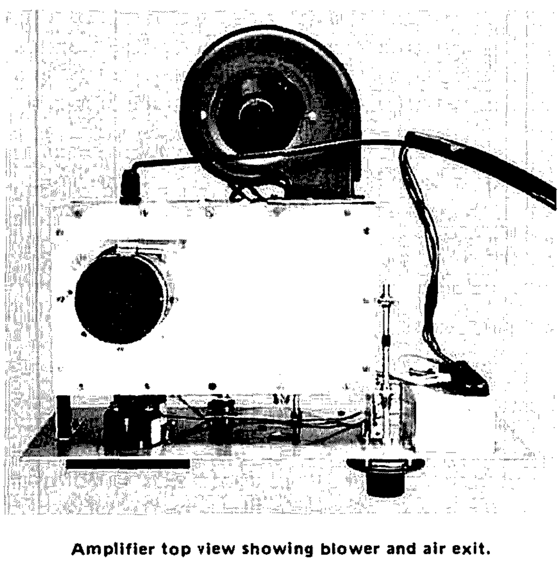
Cathode bias is connected through a feedthrough bypass capacitor and rf choke to the low rf point of the line, which is 1 inch (25.5mm) from the tube socket. The rf choke, RFC2, is made of 5½ turns of number 18 AWG (1mm) enameled copper wire. Diameter of the choke is 3/8 inch (9.5mm). Two large ferrite beads were slipped over RFC2 as added insurance against vhf parasitics.

The tuning capacitor, C1, is a 1-inch-diameter (25.5mm) brass disc mounted on the end of a surplus piston tuner. The tuner has finger-stock contacts to ensure good rf grounding of the capacitor rotating section. The stator is the end of the cathode line. Capacitor C1 is actuated by a panel-mounted knob through a long flexible shaft coupling.

The coupling capacitor, C2, is a 1-inch-diameter (25.5mm) disc of 1/8-inch-thick (3mm) double-clad PC board, which is soldered on both sides to the center conductor of a section of ½-inch-diameter (12.5mm) foam-filled, semirigid coaxial transmission line. The inner conductor of the hard line projects 3/8 inch (9.5mm) from the outer conductor of the hard line. The hard line passes into the cathode chamber through a flange mount, which is made of 5/8-inch-diameter (16mm), 0.049-inch (1.2mm) wall copper tubing. The tubing is soldered to a brass disc 1/8 inch (3mm) thick, 2 inches (51mm) in diameter. Saw cuts in the end of the copper tube allow the hard line to be clamped into position with a small hose clamp. A second clamp on the hard line prevents the transmission line from sliding in too far and closing the capacitor completely. The coupling capacitor is adjusted for minimum vswr on the drive line with the cathode circuit at resonance.



Cathode-tuning box with cover removed (upper left). Coils are rf chokes, which are described in the text.



Amplifier top view showing blower and air exit.

Tube socket. The tube socket was home brewed from a surplus 2C39 socket. Looking down into the plate compartment, the outer ring of the tube socket is the grounded-grid ring. It is made of Instrument Specialties type 97-135 finger stock. The finger stock is soldered into a 2-3/16-inch-diameter (51.5mm) hole in a brass sheet, which is 1/8 inch (3mm) thick and 3-5/8 inches (93mm) in diameter.

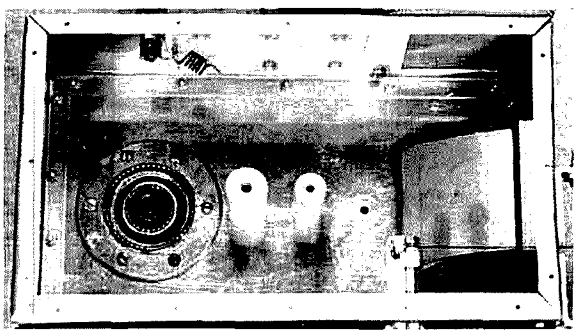
The grid ring assembly is bolted to the plate-compartment base. The grid ring mates over a 1-3/4-inch-diameter (45mm) hole in the base plate. The remaining rings in the tube socket (moving inward in

order) are the cathode ring, outer filament ring, and center cathode pin collet. The central rings are supported by the tube socket, which is mounted between two layers of ¼-inch-thick (6.5mm) plexiglass. These plexiglass layers insulate cathode and filament from ground.

The 2C39 tube socket adapted for this project has a coaxial set of brass tubes mutually insulated by a mylar sleeve. The outer tube is connected to the cathode ring and cathode line. The inner tube is connected to the outer filament ring. The center pin passes through a fused-glass bead up through the center of the socket. The center pin terminates in the filament-pin collet.

The cathode ring is a 2C39 plate ring (Instrument Specialties part no. 90-70), which is soldered inside a 1-3/8-inch (35mm) OD length of copper tubing. The tubing is shimmed to fit with copper flashing material. The filament ring is made from a short length of 5/8-inch (16mm) OD copper tubing, 0.049 inch (1.2mm) wall, which was slotted with a hacksaw. The tube filament socket fits over the filament ring.

A 4½-turn close-spaced coil 7/8 inch (22mm) in diameter connects to the filament ring, and a 5-turn close-spaced coil, ½ inch (12.5mm) in diameter, connects filament voltage to the center pin. Both coils are made of number 10 AWG (2.6mm) enameled copper wire, and both coils are connected to 0.001-µF feed-through bypass capacitors mounted on the compartment side wall.



**Amplifier showing plate-circuit components but with plate line and tube removed. Note tube socket and bypass capacitor arrangement.**

Blower requirements. The blower is mounted at the plate-tuning end of the plate-circuit compartment. The air-inlet hole is covered with aluminum insect screen. Air is forced into the plate-circuit compartment by the blower and is restricted to exiting through the anode fins by the plate line and a mylar chimney, which is connected between the anode outer diameter and the top-plate air exit. A Dayton type 2C610 blower is used.

Measurement of the differential pressure across the tube anode, with all air-system components in place, indicated ¼ inch (0.5mm) water-pressure drop. According to the Eimac data sheet for the 8938 tube, this pressure drop corresponds to an air flow of 28 cfm (0.79cmm), which is sufficient for safe operation at 1-kW plate dissipation at sea level.

The mylar chimney is 3½ inches (89mm) in diameter – and 2½ inches (64mm) long. It is formed from 5-mil (0.03mm) mylar. I found that mounting the plate-circuit

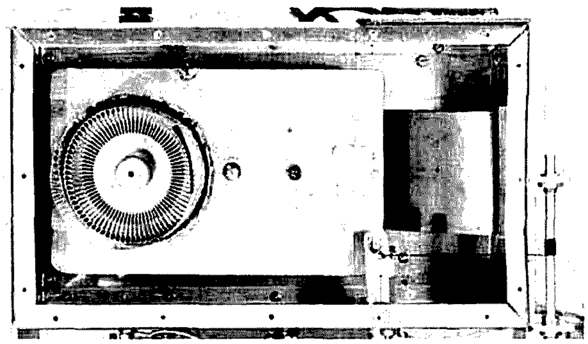


**Plate-circuit compartment showing plate line, mylar chimney, and air exit.**

finger stock upside down created a convenient cup for one end of the mylar chimney.

The tube was installed into the socket first, then the plate line was pushed down over the tube anode and the plate line was screwed onto the ceramic standoff insulators. The mylar chimney was then placed over the tube anode and carefully fitted into the finger stock cup. The top plate was aligned with the chimney and screwed down, using self-tapping screws. Blind fasteners can be used for an rf tight seal. Remember to solder the plate rf choke to the plate line before securing the cover!

The top cover is a 1/16-inch-thick (1.5mm) aluminum sheet, 1 inch (25.5mm) wide, which is fitted with an air outlet assembly. This assembly is made from a 1/16-inch-thick (1.5mm), 1-inch-wide (25mm) sheet-metal strip, which was bent around the tube anode for sizing, then cut. Several fingers were cut into the strip with a fine-blade saw, then each finger was bent up.

A ½-inch-thick (12.5mm) piece of aluminum honeycomb cut into a disc fits into the outlet assembly. A large stainless-steel hose clamp secures the honeycomb disc inside the outlet assembly. The outlet assembly is clamped to the top cover by a 4½ inch (115mm) square sheet of 1/16-inch-thick (1.5mm) aluminum with a 3-3/8-inch-diameter (86mm) hole cut in the center. The mylar chimney fits inside the air outlet, forming an effective air- and rf-tight enclosure.

## power supply requirements

Two-kW PEP input requires a supply nominally rated at 2500 volts at one ampere. The actual plate current at 2 kW input will depend on the value of the zener diode used for cathode bias. A 27-V zener (1N2822B) was used here. With 2200 volts, the amplifier idles at about 50-70 mA and whistles up to 800 mA on SSB or key-down CW tune position. The 1760 watts input yields over a kW output. Operation is just on the "A" side of class B. On-the-air reports are of good quality SSB.

The tube manufacturer recommends 4.0 Vac on the filament for 432-MHz operation. This can be supplied by a 5-V transformer with *Variac* adjustment of the primary voltage.

The power supply used for the South American operation was built by W3HQT. It uses *Variac* control on both the B+ and filament supplies. It is detailed schematically in **fig. 2**. The B+ primary power circuit is circuit-breaker protected. The blower comes on with the filament switch.

## bias and metering

The bias and metering circuits are pretty standard for grounded-grid amplifiers.[2] A 10k, 25-watt resistor in the cathode bias lead provides cutoff bias during standby periods. In the operate mode, this resistor is shorted by a set of contacts tripped by the ssb vox circuit. Operating bias is provided by the 27-volt zener connected in series with the cathode and B−.

The grid current is measured by a 0-100 mA meter in the cathode-to-ground (grid) lead. Plate current is measured in the B− lead by a 0-2 A meter. Two 0-200 μA API meters were used with appropriate shunt resistors. A 200-ohm, 10-watt resistor provides B− reference to ground should the grid meter open up. A second 200 ohm, 10-watt resistor prevents the cathode from rising to a high potential in the event the zener burns open.

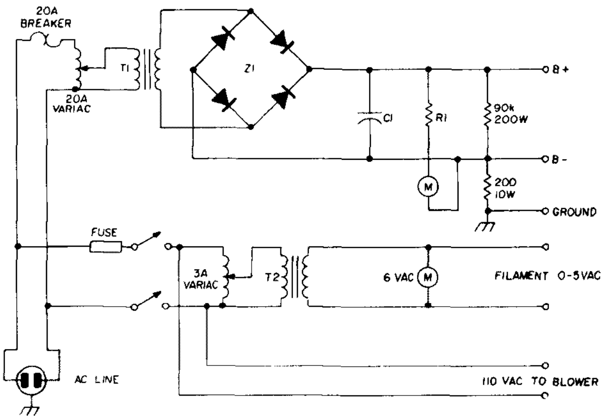



**fig. 2. Power supply for the 432-MHz rf power amplifier. The 10 megohm resistor for the B+ meter is made from ten series-connected 1 megohm, 2 watt resistors.**

## B+ bypass capacitor

The B+ lead to the plate compartment is bypassed by a homemade capacitor consisting of a 0.015-inch-thick (0.38mm) piece of Teflon sheet sandwiched between a single-sided piece of PC board and the compartment wall. The PC board is 3-1/2 inches x 4-1/2 inches (89x115mm) with a flat ground lug soldered to the copper foil at one end. This lug is connected to the high-voltage connector. The foil side of the board is in contact with the teflon dielectric, which overlaps the PC board by 1/4 inch (6.5mm) all around. Six nylon screws bolt the PC board to the compartment side wall.

The plate rf choke is 5 turns of number 16 AWG (1.3mm) enameled copper wire connected between the high voltage connector and soldered to the plate line at the low voltage point, which is located approximately at the inboard edge of the tube anode.

## operation and adjustment

Never operate the 8938 tube with rf drive but without B+ voltage. In operation, both the B+ and filament-voltage *Variacs* are turned to zero before turning on the power switches. The filament switch turns on the blower, after which the filament voltage is increased to 4.0 volts with its *Variac*.

Turn on the B+ supply and set the voltage to 1000 volts. Apply operating bias by shorting the standby resistor and observe idling current level (it should be about 35 mA with 1000 V B+ and 50-70 mA with 2000 V). Apply about 25-50 watts of drive and resonate the plate and cathode tuning capacitors, C1 and C2.

Load the amplifier by adjusting output-coupling capacitor C4 a small increment at a time while re-resonating the tuning capacitor each time the load capacitor is adjusted. Adjust for maximum output. Now adjust the rf-drive coupling capacitor, C2, in small increments and re-tune the cathode tuning capacitor each time until minimum vswr occurs on the drive line. Minimum vswr should easily be below 1.5:1.

With all controls adjusted, increase B+ to the operating level and readjust all controls for the new operating conditions. Adjust C3 and C4 for maximum output. Adjust C2 for minimum input vswr when C2 is tuned for maximum grid current.

When shutting down, remove B+ and turn down filament voltage to zero. After a few minutes cooling-off time, the blower may be switched off.

## performance

Two-kW PEP input is achieved at 60% plate efficiency with about 13 dB power gain as a minimum. Since the calibration of power meters, even those of good repute, is often in doubt, this performance figure is conservative.

No difficulties were experienced with this amplifier during the South America EME expedition. It ran reliably and well during long days of moonbounce schedules under adverse line voltage and temperature conditions.

## acknowledgement

I wish to thank Walt Bohlman, K3BPP, Bill Olson, W3HQT, and Bill Wenner, W3IVL for their help with the design and checkout of the amplifier, and Dick Knadle, K2RIW, who reviewed the manuscript.

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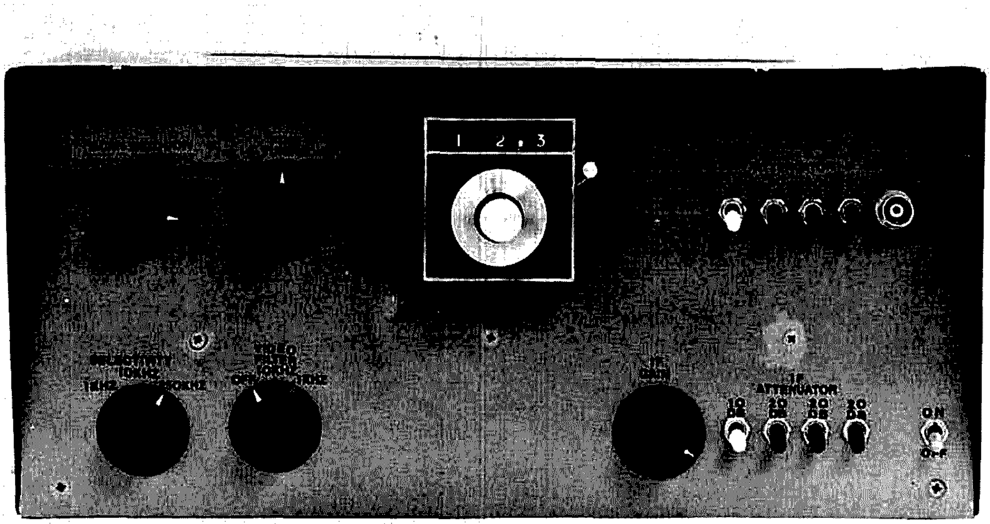
1. R. T. Knadle, K2RIW, "A Strip-Line Kilowatt Amplifier for 432 MHz," *QST*, April and May, 1972.
2. R. I. Southerland, W6UOV, "High-Performance 144-MHz Power Amplifier," *ham radio*, August, 1971.

**ham radio**



# high-performance spectrum analyzer

These instruments have been limited mainly to professional labs — now you can build your own spectrum analyzer from the information furnished here

A spectrum analyzer is a radio receiver with a swept local oscillator that allows continuous tuning over a specified frequency range. The received signals are displayed on a conventional oscilloscope as pips. **Fig. 1** shows the radio spectrum between 100 kHz and 100 MHz in the San Francisco area as displayed on the spectrum analyzer described in this article.

Because of their cost and complexity, good spectrum analyzers have been limited primarily to research and development activity. Some military instruments have shown up on the surplus market, but poor sensitivity and selectivity, images, spurious responses, and poor dynamic range have limited their usefulness.

Considerable time was spent in the design of the spectrum analyzer described here. Even more time was required to assure circuit reproducibility, minimize the variety of parts, and select the least-expensive components. No PC boards are used in the design nor are any planned. PC boards in the rf and i-f strips, unless very carefully made, would probably degrade the performance of the instrument. Complete design, construction, and testing details are included for those wishing to build the spectrum analyzer.

## applications

The spectrum analyzer can be used to observe harmonics, parasitic oscillations, and sidebands of CW, a-m, fm or ssb signals. Propagation conditions in terms

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of station activity can be observed by looking at the number of signals on an amateur band. This suggests another use for DX and contest operating: You can immediately see where the pileups are without cranking the receiver tuning dial back and forth. The spectrum analyzer described in these pages was built to look for the source of birdies in an experimental general-coverage radio receiver.

The conventional way to observe radio signals is in the time domain on an oscilloscope. This is a good method when no harmonics, parasitic oscillations, or other signals are present. However, amplifiers can oscillate, modulators or mixers might not reject signals being modulated, oscillators may be oscillating on more than one frequency, or detectors may be passing the signal being detected. These signals might appear on an oscilloscope in the time domain as confusing pictures, as shown in fig. 2A. Fig. 2B shows a signal and behind it the second harmonic. This is the frequency domain. If a signal fundamental is on 5 MHz and the second harmonic in on 10 MHz, these appear on a spectrum analyzer as pips at 5 and 10 MHz. Instead of seeing a complex waveform in the time domain as on an oscilloscope, the signal is viewed in the frequency domain. The amplitude and frequency of each component can be seen (fig. 2C).

## three kinds of spectrum analyzers

Fig. 3 shows a real-time analyzer. The incoming signal is connected to a series of filters followed by detectors. A scan or sweep generator drives the horizontal plates of an oscilloscope and controls an electronic switch, which selects the proper detector output. The frequency range is limited by the number of filters and their bandwidth. This type of analyzer is quite expensive because of the large number of filters required to cover a spectrum. Its main application is in the audible or subaudible range.

Another type is the tuned radio-frequency analyzer shown in fig. 4. The input filter is a tunable bandpass filter, which is tuned by the scan generator. The TRF analyzer lacks resolution and sensitivity. Also, tunable filters usually don't have constant bandwidth. The TRF analyzer is used mainly for microwave analysis. There are other types of analyzers, such as a digital processor that performs Fourier analysis, but these are very complex.

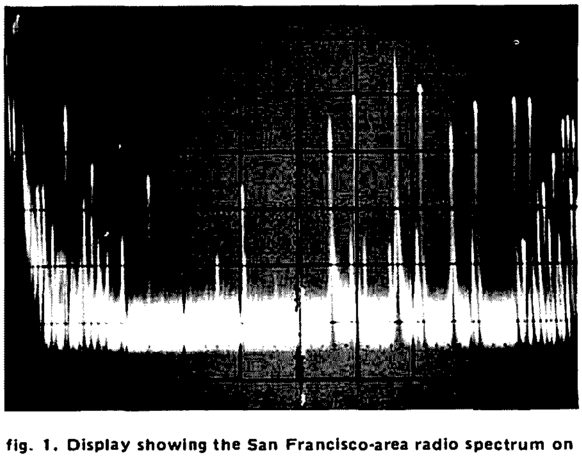


fig. 1. Display showing the San Francisco-area radio spectrum on the spectrum analyzer. A TV antenna was used to receive the signals on the right half of the display, and the TV antenna lead was used as a long-wire antenna to receive signals on the left half.

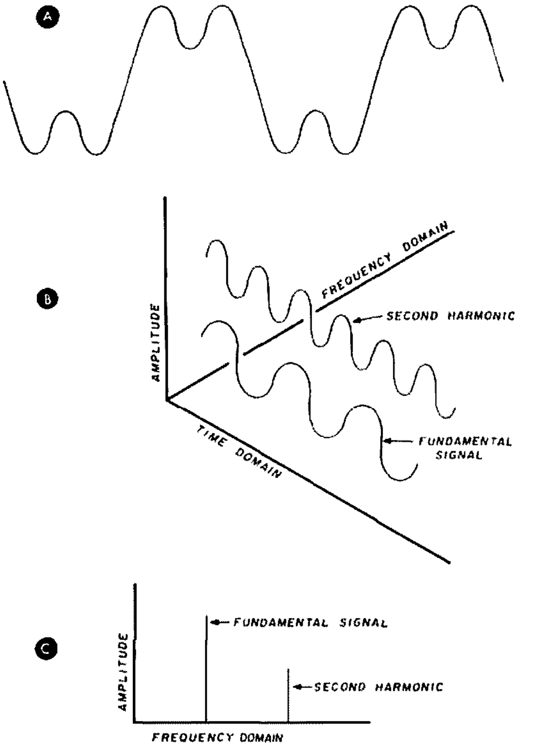



fig. 2. A fundamental-frequency signal and its second harmonic as shown on an oscilloscope in the time domain, A. The same signal and its second harmonic are shown in B in the frequency domain to illustrate their relationships. The spectrum analyzer displays a fundamental signal and its second harmonic in the frequency domain as shown in C. Amplitude and frequency of each can be seen.

The most common type of spectrum analyzer is the superheterodyne, as shown in fig. 5. It is similar to a standard superheterodyne receiver with the following exceptions: There is no tuned circuit in the front end, because it would be difficult to make one continuously tunable from 100 kHz to 100 MHz, and the local oscillator is electronically tuned by the scan generator from the oscilloscope. The superheterodyne analyzer has the advantage of a wide tuning range and controlled bandwidth. Also the sensitivity can be made uniform. Image problems are minimized by triple conversion.

Fig. 6 shows the spectrum analyzer described here. No more than 1 volt rms should be applied to the input attenuator. The output from the attenuator should be limited to about 10 mV rms maximum. The function of each assembly is discussed with reference to the block diagram. Later, theory of operation, circuit design,

construction, and checkout are described for each assembly.

The input attenuator is followed by an LC 130 MHz lowpass filter. Its purpose is to prevent incoming signal harmonics from mixing with the first local oscillator and producing spurious responses. The incoming signal, 0.1 - 100 MHz, is mixed with the vco to produce a 200-MHz first i-f. The vco is driven by a sweep shaper that predistorts the sawtooth wave from the oscilloscope to compensate for tuning nonlinearities of the varactor tuning diode in the vco.

The 200-MHz i-f output from the first mixer (all mixers are double balanced) goes to an amplifier, which compensates for the mixer loss. Next is a 200-MHz bandpass filter followed by the second mixer. The 200-MHz i-f signal mixes with the second local oscillator output to provide a 50-MHz i-f signal. This signal is amplified and applied to a 50-MHz bandpass filter. The output of this bandpass filter is mixed with 39.3 MHz to produce a third i-f of 10.7 MHz. The 10.7 MHz output from the third mixer is then amplified and fed through a 250-kHz bandpass ceramic filter. This stage is followed by an i-f attenuator.

Next are two identical crystal filters to provide 1- and 10-kHz *bandwidths. These feed a gain equalizer, which* compensates for the differences in gain between the various bandwidths. Completing the i-f section is an amplifier that drives a second 250-kHz bandpass ceramic filter.

The logarithmic amplifier compresses the output signal so that a large-amplitude signal can be displayed on an oscilloscope. Two divisions on the scope correspond to a change of 10:1, or 20 dB. Six divisions represent a change of 1000:1, or 60 dB. Without the log amp, a CRT about 30 inches (76cm) tall would be required to display the same information with the sensitivity available at the top of the display. A video filter is included to remove high-frequency noise in narrow-bandwidth operation. The performance specifications for the spectrum analyzer are listed in **table 1**.

**table 1. Performance specifications of the W6URH spectrum analyzer.**

| | |
|---|---|
| **frequency range:** | 0.1 MHz - 100 MHz. |
| **sensitivity:** | 10 $\mu$V or less in the 1-kHz position. |
| **selectivity:** | Approximately 1, 10, or 250 kHz (switchable). |
| **sweep width:** | 0, 5, 10, 100, 500, 1000, and 10,000 division. In the 10,000 kHz or 10 MHz/division position, the center frequency is set to 50 MHz. Otherwise, frequency is controlled by the center control on the front panel. |
| **first LO frequency jitter:** ≈ 5 kHz | |
| **signal separation:** 10 kHz for 40-dB resolution in 1-kHz position. | |
| **input impedance:** 50 ohms | |
| **response 100 kHz - 100 MHz:** ±2 dB | |
| **frequency accuracy:** ±3 MHz. | |
| **video filter:** 10 and 1 kHz. | |
| **shape factor 3 - 60 dB:** | 250 kHz 2.5:1<br>10 kHz 50:1<br>1 kHz 30:1 |

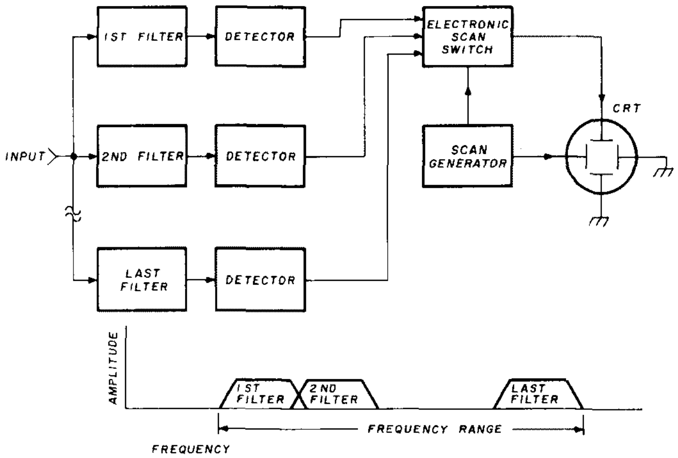



**fig. 3. A real-time spectrum analyzer. Filters are spaced to provide continuous coverage of the frequency spectrum.**

The display scope for the spectrum analyzer should be dc coupled, have 0.1 V/division vertical sensitivity, and have six divisions of vertical calibration by ten divisions of horizontal calibration. Also approximately 6V of horizontal sawtooth voltage should be available to drive the spectrum analyzer from the scope. Scope input impedance should be 1 megohm and input capacitance 10 - 40 pF.

## circuit description

This section presents the theory of operation and design details of each of the major circuits in the spectrum analyzer. It is recommended that the following text be read before attempting to build the circuit modules as much information is provided that will be helpful when construction is started. Especially important are the suggestions on decoupling, shielding, and bypassing.

**Sweep shaper.** The sweep-shaper schematic is shown in **fig. 7**. The resistors in the input circuit, which are switched by S601A, determine the sweep width. In the 10-MHz/division position, the sweep center frequency is set to 50 MHz by R101. In all other positions the sweep center frequency is set by R601. R104 sets the low-frequency limit, 0.1 MHz, and R102 sets the high-frequency limit, 100 MHz. R103 is a fine frequency adjustment with about 500 kHz of range. U101 provides isolation for the input attenuator and some gain. U102 provides isolation for the frequency control. The outputs of U101 and U102 are resistively combined and drive U103, which is a nonlinear amplifier that complements the nonlinearities of the tuning diode, CR203, in the vco.

As the frequency is increased, more voltage is required. If a change of 1 volt is required to go from 260 to 270 MHz, a change of 1.3 volts may be required to go from 270 to 280 MHz. **Fig. 8** gives an idea of the distorted waveform available from U103 and the actual waveform required to compensate for nonlinearities of the tuning diode in the vco. Compensation is usually required between 270 and 300 MHz. When the voltage

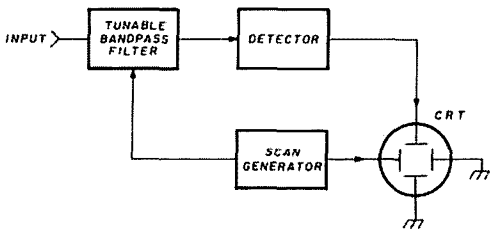


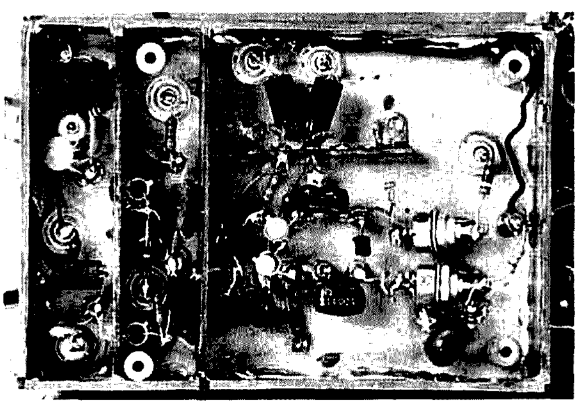

**fig. 4. Tuned radio frequency (TRF) spectrum analyzer. Tuning is accomplished by adjusting the bandpass-filter frequency.**

on pin 3 of U103 reaches the voltage on the right side of R115, gain will be increased by the setting of R115, and so on up the line, to R106. R115 sets the nonlinearity at 10 MHz and R106 at 100 MHz. The tuning diode used in the vco I built was linear up to about 280 MHz.

Care must be taken so that no low-frequency components, including power-supply hum, are added to the incoming sawtooth wave by the sweep shaper. Added low-frequency components will result in first local-oscillator instability. This shows up as jitter in the display when looking at narrow scan widths.

Voltage-controlled oscillator. Q201 and Q202 together with L201, L202, and CR203 form a 200-300 MHz voltage-controlled oscillator (vco — fig. 9). Q203 is a buffer amplifier. Q204 and Q205 provide outputs to the first mixer and a second output.

Several compromises were made in the design of the vco. To achieve high frequency stability, the oscillator should have a high C/L ratio; however, to tune it with a varactor, a low C/L ratio is required. Varactor circuits with reasonably high Q at these frequencies have a relatively small tuning range. When considering tuning range, linearity, and Q the Motorola Epicap tuning diodes seem to be about the best. Careful design of the oscillator and amplifiers was necessary to provide a constant output level to the first mixer. The 2N2369A did not have sufficient gain-bandwidth product to provide constant output when used as an oscillator transistor, but because of the low gain requirements of the amplifiers, the 2N2369A is satisfactory in these circuits.



**Closeup view of the voltage-controlled oscillator. Main circuit is in the compartment to the right; connectors for output to the first mixer and tracking generator are in compartment on left.**

Rf section. The rf section (fig. 10) contains the first two i-f stages operating at 200 and 50 MHz. The third i-f, operating at 10.7 MHz, is contained in the i-f section.

The input signal is applied to the first mixer, MX301, through an rf attenuator and a lowpass filter. The attenuator is required to keep the input to the first mixer below approximately 10 mV. MX301 combines the 0.1 - 100 MHz input with the 200 - 300 MHz signal from the vco to produce the first i-f of 200 MHz. Q301 provides gain to compensate for the loss through MX301. L305 through L307 form a 200-MHz bandpass filter. Output from the 150-MHz local oscillator, Q305, mixes with the 200-MHz signal in MX302 to produce a 50-MHz i-f. The 50-MHz signal is amplified by Q302 and Q304 then applied to the third mixer, MX303, through a four-section bandpass filter. The signal is then mixed with the 39.3-MHz output from Q306 to produce 10.7 MHz.

Several steps were taken to minimize spurious and image responses. The input to the pass filter attenuates harmonics that would otherwise mix with the first local oscillator signal, producing spurious signals. Using a first i-f twice, the highest input frequency minimizes basic image problems. Intermediate frequencies separated 5:1

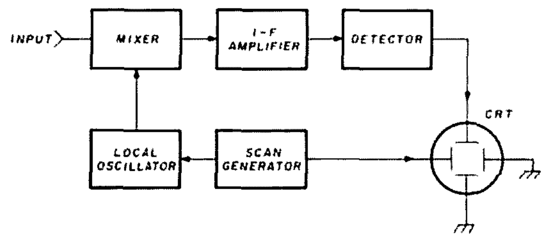



**fig. 5. Simplified block diagram of a superheterodyne spectrum analyzer. Local oscillator is tuned by the scan generator.**

or less from the next i-f further reduce any images. Double-balanced, four-diode mixers produce the fewest unwanted products and have relatively high local-oscillator isolation. Shielding and bypassing prevent the three local oscillators and their harmonics from mixing and producing spurious signals. For example, with inadequate shielding, the 39.3-MHz local oscillator signal would leak back through the front end. A local signal on 6 meters would be picked up by the second i-f (50 MHz). Ideally each section should be in a die-cast box with an rf gasket, but the construction with copper-clad board, described later, is adequate if care is taken to ensure an rf-tight enclosure.

A ferrite bead is used on all dc leads on each side of the feedthrough. Piston trimmers are required for the 200-MHz bandpass filter for mechanical rigidity. Links L309 and L311 should be constructed so they can be moved from 0 to ½ inch (0 to 12.5mm) away from their respective coils.

I-f section. The third-mixer output, 10.7 MHz, is amplified by Q401 (fig. 11). Q402 provides drive at 300 ohms to FL401. Next, an emitter-follower, Q403, provides drive at 50 ohms to the i-f attenuator. The two crystal filters are identical, and the first, Y401, is described.

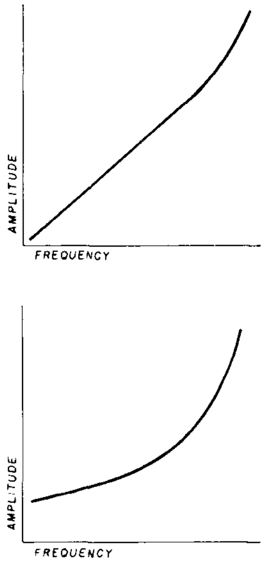



**fig. 8. Top curve shows wave shape available from U103 in the sweep shaper; bottom curve shows actual wave shape required to compensate for the nonlinearity of the tuning diode, CR103, in the vco.**

Q404 provides a paraphase output, which drives Y401 and neutralizes its capacitance. Q405 provides a high input impedance to the crystal-filter output. CR401 provides a bypass around Y401 for 250-kHz bandwidth. CR402 switches in R401 to broaden the crystal filter for 10-kHz bandwidth. R402 adjusts the gain in the 1-kHz bandwidth position, and R403 adjusts gain in the 10-kHz bandwidth position to the same as in the 250-kHz bandwidth positions. CR403 switches in R403 in the 10-kHz bandwidth position. Q408 and Q409 provide a gain of about 40. Q409 provides 300-ohm drive to FL501 in the log-amp section.

FL401 and FL501 provide 250-kHz bandwidth. After manufacture, these filters are separated into groups and are color coded as to their actual center frequency, which are around 10.7 MHz. Because of spurious resonances above the crystal-filter frequencies, *orange and only orange* color-coded filters maybe used. With the orange-colored ceramic filters, crystal-filter spurious responses are down −50 to −60 dB. With opposite-end ceramic filters, they may be down only 15 dB.

One of the more challenging aspects of designing a spectrum analyzer is selectivity. In its widest position, the spectrum analyzer should have a bandwidth to observe signals using a sweep width of 100 MHz. In its narrowest position, it should be able to resolve signals close together, such as a carrier and its sidebands. Spectrum analyzers with elaborate sets of carefully matched crystal filters can achieve bandwidths as low as 10 Hz at the 3-dB points. These crystals are especially ground to

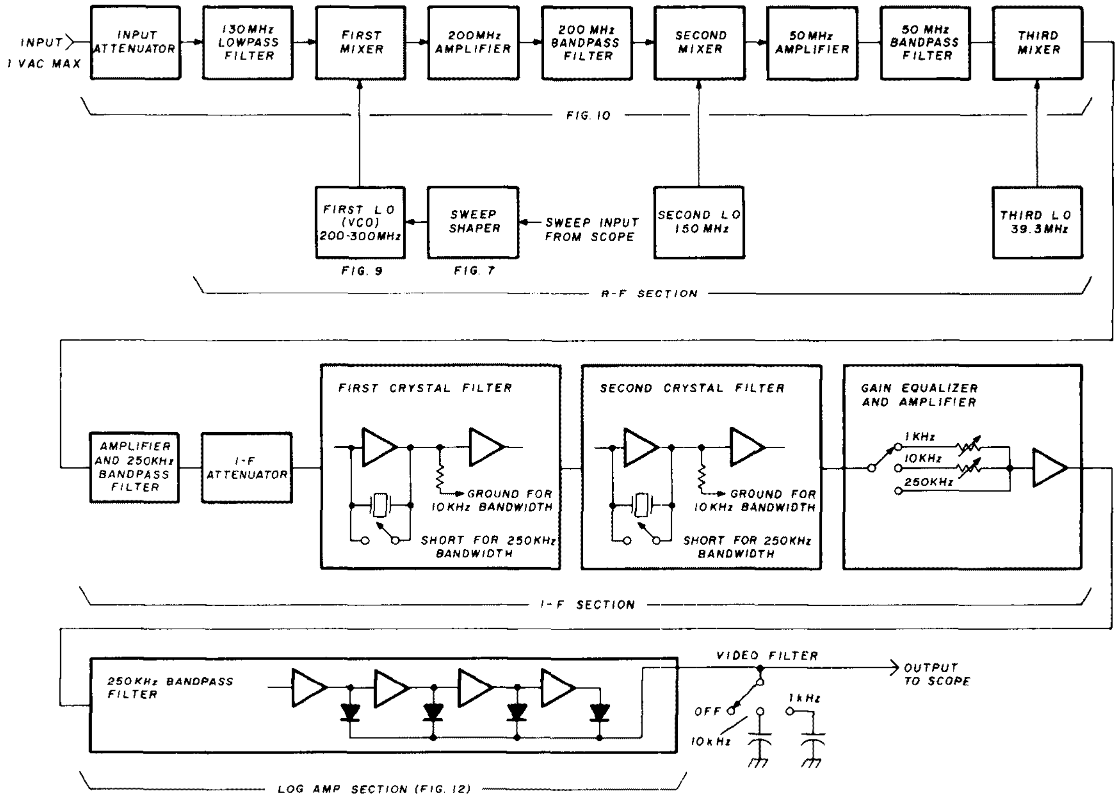



**fig. 6. Complete superheterodyne spectrum analyzer described here. Each block represents an individually shielded section or assembly.**

minimize resonances other than those desired for crystal filters.

Another important consideration is shape factor and ringing. Communications receivers often have a 3-60 dB shape factor of 2:1. Unfortunately, these filters with steep skirts have phase discontinuities at their band edges, and therefore ringing occurs when signals are rapidly swept through them. This phenomenon was demonstrated using a crystal filter from an fm transceiver in the spectrum analyzer. Shape factor of narrow-band filters used in spectrum analyzers cannot be as low as that found in communications receivers.

The ceramic filter used in the 250-kHz bandwidth position has a shape factor of about 2.5:1. These filters are used mainly in fm automobile radios.

The crystal filters were made using off-the-shelf crystals from the local Heathkit store. The filters exhibit some resonance, −50 to −60 dB, slightly above the filter frequency. With a shape factor of about 30:1 two signals about 10-kHz apart have about 40 dB of resolution. The filters are not sufficiently narrow to look at 1- or 2-kHz sidebands of a transmitter. To observe signals which are separated by only 1 kHz would require a 500-Hz filter with a shape factor of 2:1, a sweep speed of 5 sec/cm, a

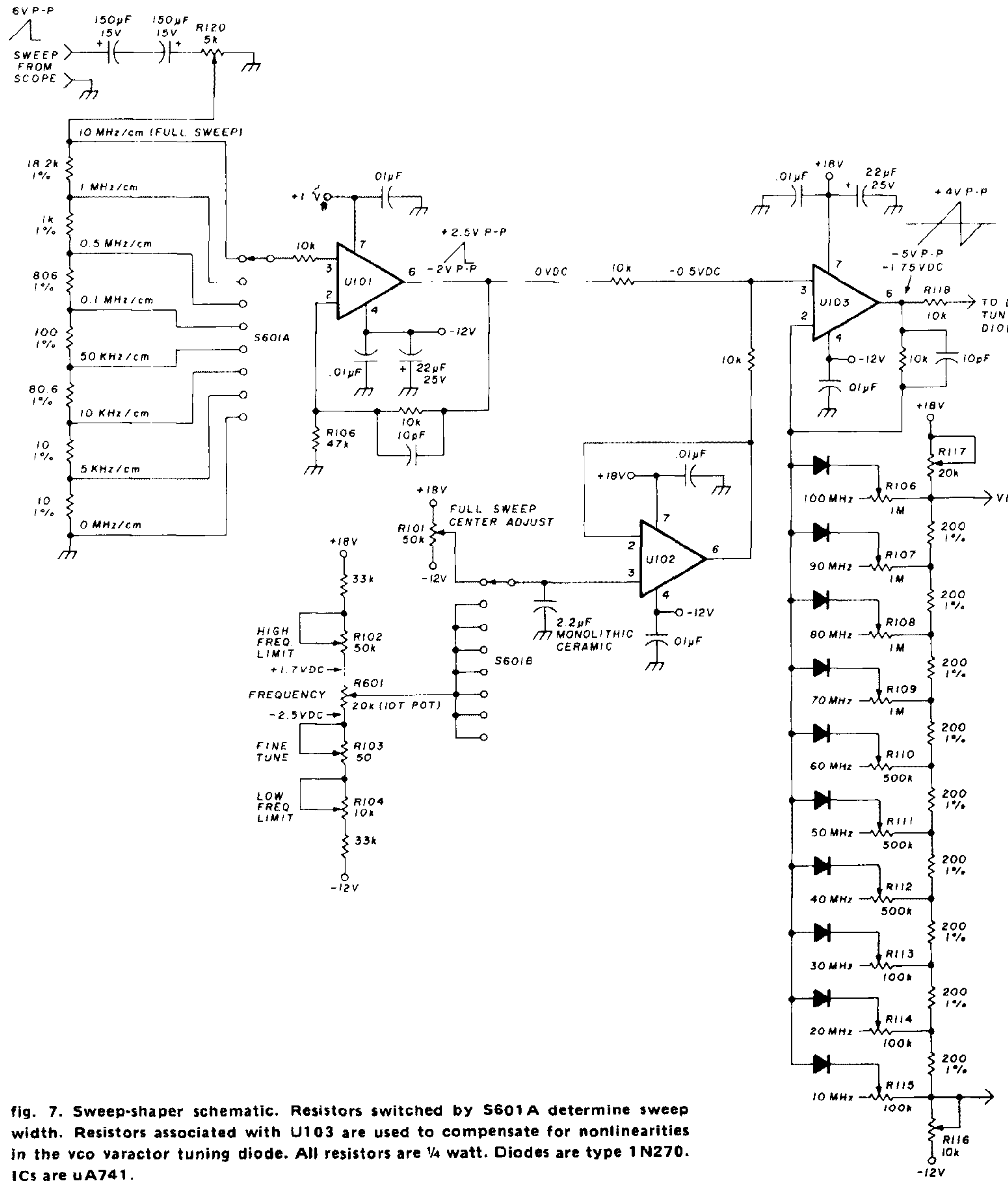



**fig. 7. Sweep-shaper schematic. Resistors switched by S601A determine sweep width. Resistors associated with U103 are used to compensate for nonlinearities in the vco varactor tuning diode. All resistors are ¼ watt. Diodes are type 1N270. ICs are uA741.**

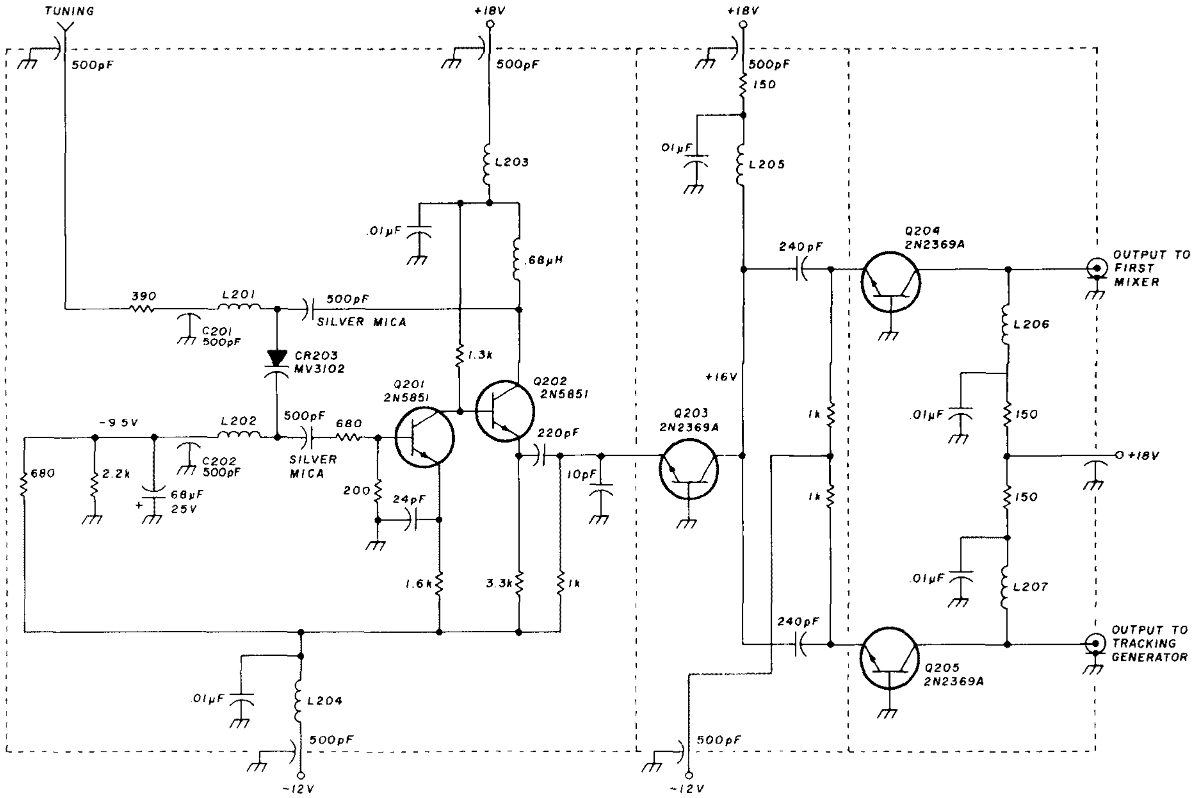



**fig. 9. VCO schematic. Range is 200-300 MHz. CR203 is a Motorola MV3102. C201 and C202 are 500-pF feedthrough bypass capacitors. L201, L202: 1 turn no. 18 AWG (1mm) wire. L203-L207: 5½-turn chokes.**

phase-locked vco to remove vco fitter, and a storage scope because of the very slow sweep speed.

**Log amplifier.** The log amp schematic is shown in **fig. 12.** FL501 provides selectivity in the 250-kHz position. Emitter-follower Q501 provides a low-impedance input for Q502. Q502 through Q505 are a series of four wideband resistance-coupled amplifiers. Each stage has a gain of 6 for a total gain of 1296, or slightly over 60 dB.

The detected output of each stage is summed through each diode at the output. As the signal level at the output of Q505 increases, CR505 conducts with a logarithmic-shaped curve. Q505 then saturates and CR504 starts conducting, and so on down the line. CR502 and CR503 are in series to provide a voltage greater than that of CR503 and CR504. R501 through R503 set the output level from each stage, and R504 sets the total output level. The video filter reduces high-frequency noise. It can be used only for narrow sweep widths. The video filter is described in detail in the section on operation.

## construction

All assemblies except the sweep shaper are built in boxes made of 1/16-inch thick (1.5mm) single-sided copper-clad board. The rf assembly has double walls, as explained later, and individual covers. All other assemblies have single-wall separators and one cover for the entire assembly. All boxes are 3/4 inch (19mm) high (inside measurement), and the covers are secured with 3/4-inch long (19mm) 4.40 (M3) metal spacers. The following parts had to be obtained from the sources shown. All other parts came from the junk box and the sources are unknown.

| component | source |
|---|---|
| ceramic filters (FL401, FL501) Vernitron FM-4 | Vernitron, 232 Forbes Road, Bedford, Ohio 44146 |
| crystals (Y401, Y402) Heathkit 404-39 | local Heathkit store |
| 2N2369A, J309 Siliconix | James Electronics, P.O. Box 882, Belmont, CA 94002 |
| one-hole beads (Amidon FB-43-101) six-hole bead (Amidon FB-64-5111) | Order from Amidon Assoc., 12203 Otsego St., North Hollywood, CA 91607 |

**Sweep shaper.** This assembly is built on copper-clad boards without shielding or feedthrough bypassing.

**vco.** The vco is built in a copper-clad box measuring 4½x3x¾ inches (114x76x19mm). The buffer section and output amplifiers are shielded from the oscillator. Paper-thin copper is wrapped over the outside edges where the cover attaches.

**rf section.** The rf section has double walls between compartments, which ensures good shielding. The walls are spaced 1/16 inch (1.5mm) apart, and paper-thin copper (available from hobby shops) is laid over the walls. The covers for the two bandpass filters are cut as shown in **fig. 13** (a continuous shield will create a

shorted turn, which will seriously detune the filters). All assemblies have 500-pF feedthrough capacitors for dc voltages and 0.01-$\mu$F capacitors across the feedthrough caps inside the compartments. Dimensions of the rf section are shown in **fig. 14**.

The construction of the three mixers is identical. The transformers for each unit use four trifilar-wound turns. All windings are wound at the same time for a total of 12 turns. **Fig. 15** shows four turns through a ferrite bead. The four turns are counted from the inside — *not* the outside. Also shown in **fig. 15** are the transformer connections and how they are installed in the mixer. All ports have baluns consisting of two bifilar-wound turns on ferrite beads. The same type bead is used for the transformers and baluns.

**I-f section**. Dimensions for this assembly are shown in **fig. 16**. It has one large cover rather than individual covers as used in the rf section.

**power supply**. The power supply is located beneath the chassis at the rear. Two separate supplies are required (**fig. 17**). Transformer T701 should provide 23-29 Vdc at approximately 200 mA and T702 should provide 18-24 Vdc at 100 mA. Both regulators are commonly available three-terminal ICs. The supplies must not go out of regulation above about 100 Vac line voltage.

## checkout

Before any tests are made, all dc voltages should be checked against those shown in the schematics. In many instances, each stage of an assembly is checked. In each case, the generator should be terminated and coupled through a 0.01-$\mu$F capacitor. The existing input to the stage should be removed during these checks. The generator should then be carefully tuned to the frequency that produces maximum deflection on the display scope. Stage-by-stage checkout is important because it is otherwise difficult to determine which stage is at fault if performance is not satisfactory. Assembly checkout is followed by a final alignment.

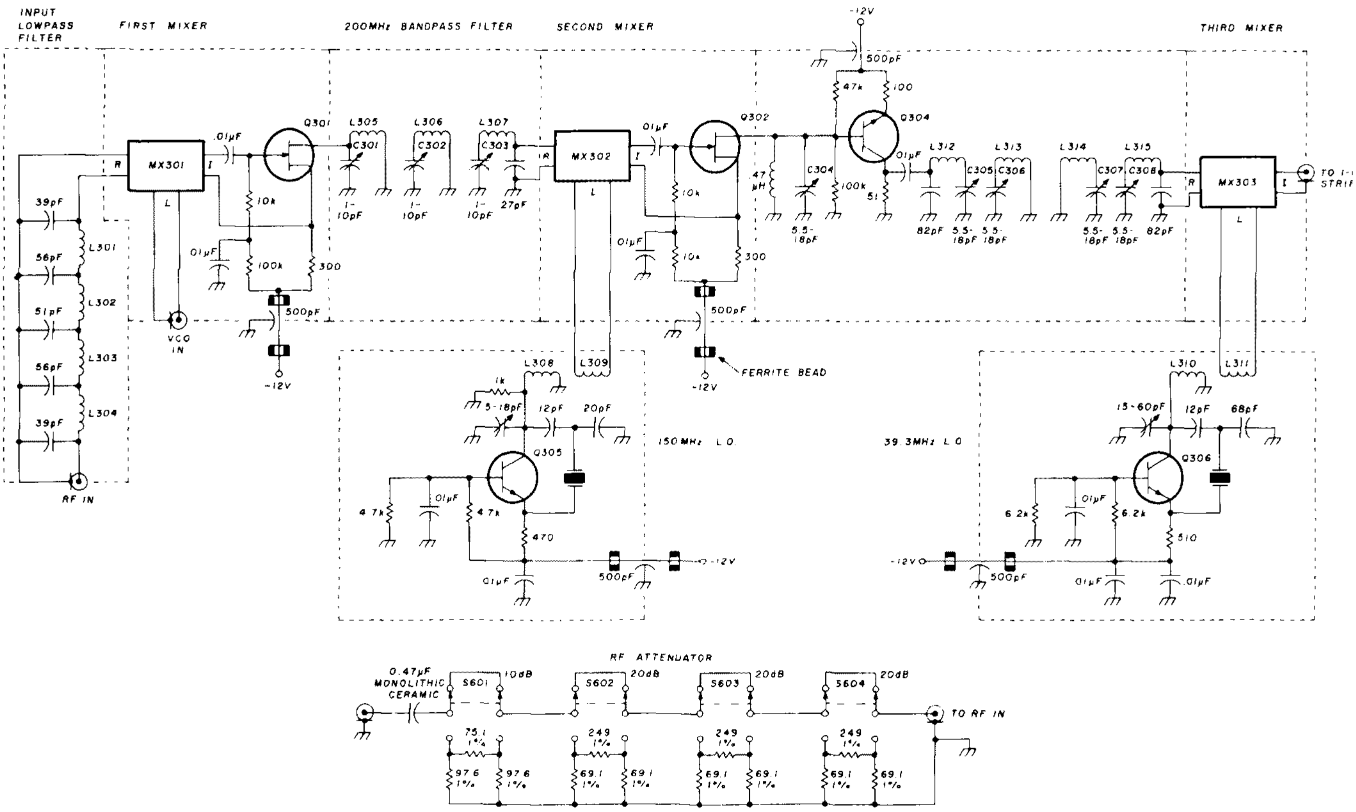



**fig. 10. Rf-section schematic. All bipolar transistors are 2N2369As; fets are J309 or J310. Feedthrough caps are 500 pF.**

| | |
|---|---|
| L301-L304 | 4 turns no. 18 AWG (1.0mm), 3/16" (5.0mm) diameter, ¼" (6.5mm) long |
| L305-L307 | 4 turns no. 22 AWG (0.6mm), ¼" (6.5mm) diameter, 3/16" (5.0mm) long. Spacing between coils ¼" (6.5mm). All wound on one ¼" (6.5mm) diameter plastic form. |
| L308 | 3 turns no. 18 AWG (1.0mm), ¼" (6.5mm) diameter, ¼" (6.5mm) long |
| L309 | 2 turns no. 22 AWG (0.6mm), ¼" (6.5mm) diameter, close wound |
| L310 | 10 turns no. 18 AWG (1.0mm), 5/16" (8.0mm) diameter, close wound |
| L311 | 2 turns no. 22 AWG (0.6mm), ¼" (6.5mm) diameter, close wound |
| L312-L315 | 7 turns no. 26 AWG (0.3mm), ¼" (6.5mm) diameter, close wound. Spacing between coils is ½" (12.5mm) |

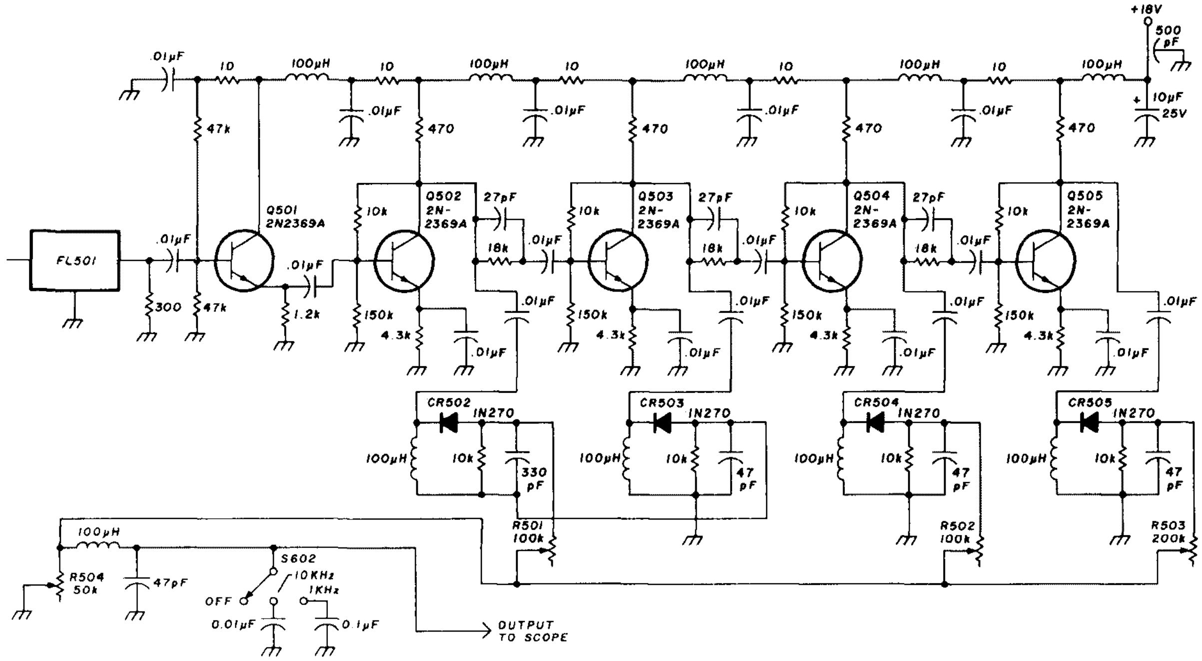



fig. 12. Log amplifier schematic.

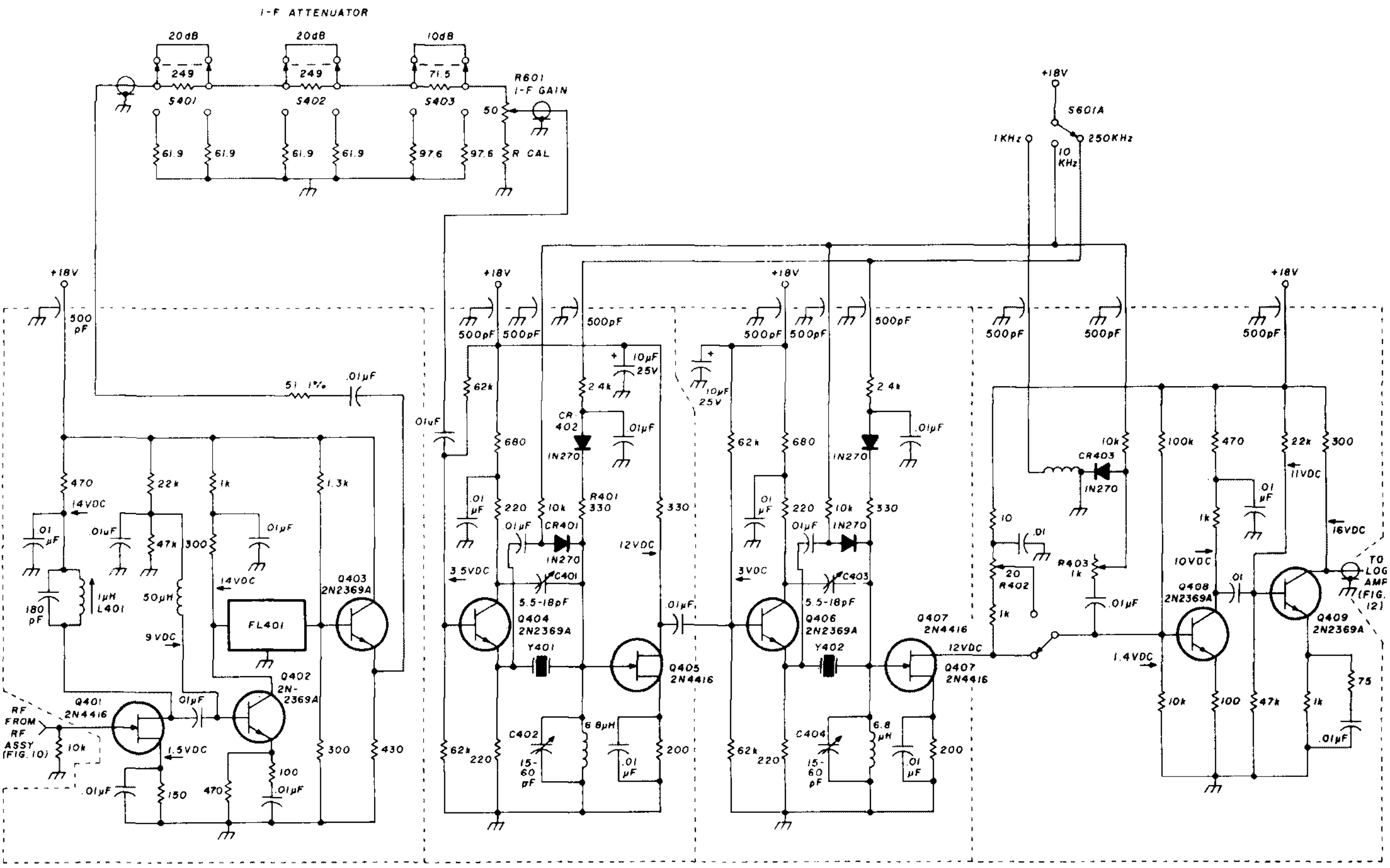



**fig. 11. I-f section schematic. All resistors in the attenuator must be 1% tolerance. Resistors must be added in series or parallel to obtain 51 ohms ±1% with R601. Feedthrough caps are 500 pF.**

### test equipment required

1. Rf signal generator, 10-300 MHz, with calibrated attenuator 1 $\mu$V to 0.1V rms (Measurement model 80 or HP 608 or equivalent).

2. Grid-dip oscillator, 10 MHz - 200 MHz (Heath GD1 or equivalent).

3. Rf signal generator, 100 kHz - 10 MHz. Calibrated attenuator 1 $\mu$V - 0.1 V desirable (General Radio 605 or equivalent).

4. Volt-ohmmeter (Triplet 630 or equivalent).

5. Comb generator. (Prescalar described in June, 1975, *ham radio*[1] not absolutely necessary but convenient for adjusting vco linearity).

6. Frequency counter capable of counting to 300 MHz.

**Sweep shaper and vco.** Verify the voltages at the output of U101 and U103 against those shown in the schematic, fig. 7. Using a 300-MHz counter, determine what tuning voltages in the vco produce 200, 250, and 300 MHz. (Further checkout of the vco and sweep shaper is described in the section on final alignment).

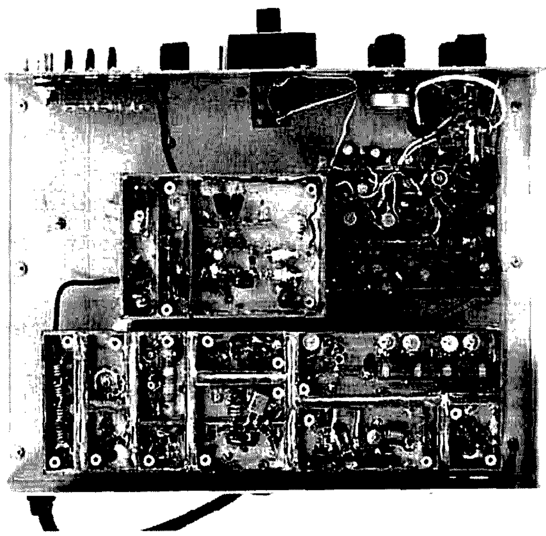
**Log amplifier.** Check dc voltages as shown in fig. 12. Ensure that 2-3 volts are between collector and emitter of Q502-Q505.

1. Set all pots to center range.

2. Connect generator to log-amp input.

3. Set display scope to 0.1 V per division.

4. Set generator for 300 $\mu$V output and tune generator for peak on display. Verify according to the chart shown in the next column.



Top view of spectrum analyzer chassis showing the rf assembly (along rear of chassis, bottom), voltage-controlled oscillator (center), and sweep shaper (top right).

| | |
|---|---|
| 300 $\mu$V | $\approx$ 1 division |
| 1,000 $\mu$V | $\approx$ 2 divisions |
| 3,000 $\mu$V | $\approx$ 3 divisions |
| 10,000 $\mu$V | $\approx$ 4 divisions |
| 30,000 $\mu$V | $\approx$ 5 divisions |
| 100,000 $\mu$V | $\approx$ 6 divisions |

5. Set R504 so that 100,000 $\mu$V equals six divisions.

Log amplifier tracking is described under final alignment.

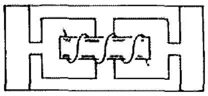


fig. 13. Foil cutout for the 200- and 50-MHz bandpass-filter covers used in the rf section.

**I-f section.** When checking dc voltages, verify voltages from S601 to the i-f section in all bandwidth positions.

1. Set bandwidth to 250 kHz.

2. Turn off i-f attenuator and turn i-f gain to maximum.

Because of noise or interfering signals, the baseline during some tests of the i-f and rf sections will shift up one or two divisions. If checkout shows that 100 $\mu$V should provide one division of deflection and the baseline is already up 1.8 divisions due to noise or external signals, this 100 $\mu$V should move the baseline up one to 2.8 divisions. Fifteen microvolts to the base of Q408 and Q404 should provide one division of deflection.

3. Connect generator to Q401 base.

4. Peak L401. Two microvolts should provide one division of deflection.

The crystal filter alignment is described in the section on final alignment.

**Rf section.** Using a grid-dip meter verify that the 39- and 150-MHz oscillators are oscillating and are crystal controlled. If the oscillators are free running, some adjustment of the two capacitors on the collector side of the crystal can be made. Turn power off and on several times to ensure that oscillators continue to run.

Perform the following steps:

1. Set bandwidth to 250 kHz and video filter to off.

2. Connect generator to input of MX303 (at lead that goes to L315).

3. Tune generator to a peak at 50 MHz and adjust L311 for maximum signal consistent with minimum coupling to L310. Ten microvolts should provide one division of deflection.

The 50-MHz bandpass filter is tuned next. The windings are critically coupled and require careful step-by-step tuning.

1. Connect generator to Q304 base and set display for about four divisions.

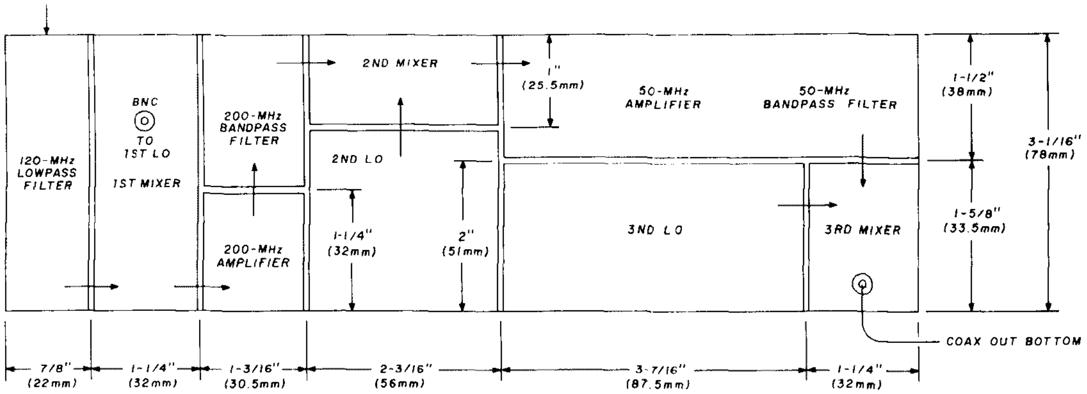



**fig. 14. Rf-section dimensions.**

2. Short L314 and adjust C306 and C308 for maximum deflection.

3. Short L313 and adjust C305 and C307 for maximum deflection.

4. Make slight adjustment of C305 through C308. Some interaction exists and several attempts will be required, but only a small adjustment will be necessary. Later, when the cover is installed, a slight readjustment will be required. Fifteen microvolts here should provide one division of deflection.

5. Connect generator to MX202 input and tune to a peak around 200 MHz. (At this frequency, signal generators sometimes drift and some readjustment may be necessary).

6. Adjust C304 for maximum deflection.

7. Set generator for exactly three divisions.

8. Install cover on 50-MHz bandpass filter and make slight adjustments of C304-C308. Twenty microvolts should produce one division of deflection.

As with the 50-MHz bandpass filter, the 200-MHz bandpass filter is critically coupled and needs careful adjustment. Proceed as follows:

1. Set all three piston trimmers to the center of their range.

2. Connect generator to Q301 gate.

3. Short L306 and peak C301 and C303.

4. Peak C302.

5. Make slight adjustments of C301-C303 for maximum deflection. Twenty microvolts here should provide one division of deflection.

6. Install covers. (Fine adjustment of the 200-MHz bandpass filter is covered under the section of final alignment).

## final alignment

This section provides final alignment information on the sweep shaper and vco, crystal filter, gain equalizer, tuned circuits, and the log amplifier. Since the sweep shaper and vco may not be familiar, some background information is given on how these circuits work together.

**Sweep shaper and vco.** The first step is to provide a sweep signal between 0.1-100 MHz. Two conditions must be met to accomplish this: The sawtooth amplitude must be adjusted so that the varactor tuning diode can tune the vco from 200 to 300 MHz. Also, the dc level must be set so that the voltage *one-half way up the sawtooth* tunes the vco to 250 MHz.

The bottom of the varactor diode, CR203, is returned to about −9 volts. The sawtooth amplitude required to tune the vco from 200 to 300 MHz is about 8 volts p-p. The voltage level required to tune the vco to 250 MHz is about −2 volts. Therefore, the sawtooth should start at −6 volts, the half-way point should be at −2 volts, and the peak should end at +2 volts. R101 sets the dc level and R120 sets the sawtooth amplitude (see **fig. 7**).

When the vco was first built, a relatively nonlinear tuning diode (1N5140) was used. It required frequency correction at most frequencies between 240 and 300 MHz to provide linear change in frequency with linear change in dc voltage. However, the MV3102 required

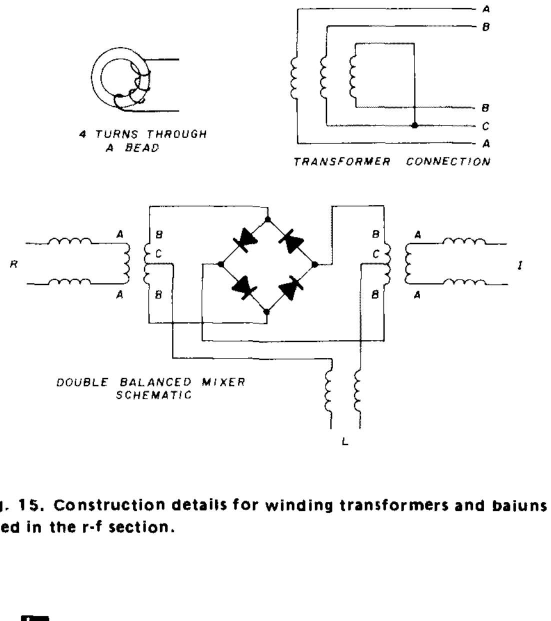



**fig. 15. Construction details for winding transformers and baluns used in the r-f section.**

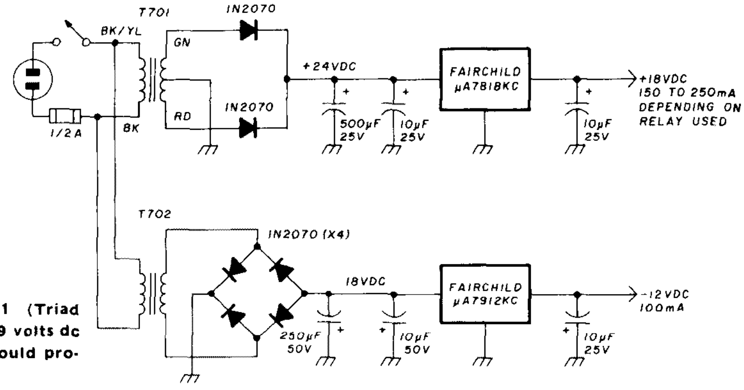



**fig. 17. Power-supply schematic. T701 (Triad F90X or equivalent) should provide 23-29 volts dc at approximately 200 mA, and T702 should provide 18-24 volts dc at 100 mA.**

correction only at 290 and 300 MHz, therefore R108 through R115 (**fig.** 7) were removed. Several MV3102s were tried with the same results; however, the entire correction circuit is included in the vco schematic.

The settings of R116 and R117 determine at what voltage, or frequency, the correction is made. If R107 is shorted, R117 can be adjusted by setting it so that a 90-MHz signal on the display is moved to the left. R106 through R115 determine the amount of correction, and R116 and R117 determine the position of correction.

A signal source that simultaneously puts out 10-100 MHz in 10-MHz steps makes vco alignment easier. The comb generator in the test equipment list fulfills this requirement. A comb generator can be made by feeding 100 MHz into the prescalar described in reference 1. The ECL logic has fast switching times and generates many harmonics, which provide the comb signal. Possibly other ECL prescalers would provide the comb signal. The frequencies generated will not be of equal amplitude.

In all scan widths other than 10 MHz/division, the center frequency is controlled by R601. R102 and R103 adjust the voltage to R601, so the oscillator is at 200 MHz when the dial reads 0 MHz and at 300 MHz when the dial reads 100 MHz. Errors of at least ±3 MHz are normal for the dial calibration.

Super heterodyne spectrum analyzers produce a signal when the local oscillator is at the i-f or 0 MHz. At 0 MHz, the local oscillator is at 200 MHz, the i-f, and energy from the local oscillator leaks through the first mixer, producing a 0-MHz marker. This is both normal and handy for calibration.

**Sweep shaper and vco alignment.** Set controls and test equipment as follows:

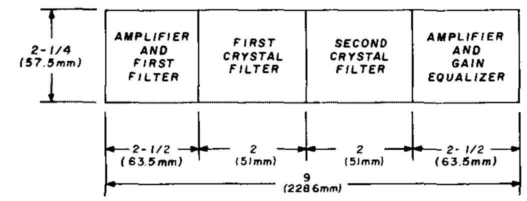



**fig. 16. I-f section dimensions.**

Bandwidth: 250 kHz.

Scan: 10 MHz/division.

Generator frequency: 50 MHz.

Generator output: ≈ 5 divisions.

Video filter: off.

1. Turn R106 through R115 to maximum resistance.

2. Adjust R101 so the 50 MHz signal appears at the center, fifth division, of the display.

3. Adjust R120 so the 0-MHz marker is at the first division. Some interaction occurs between R101 and R120. Readjust them so that the 0-MHz marker is at the first division and the 50 MHz signal is at the fifth division.

4. Look at the sawtooth with a scope and determine the dc value of the sawtooth at 0 MHz. Set R116 so this dc voltage is present on R115.

5. Look again at the sawtooth at 100 MHz and similarly set R117. Because of interaction, R116 and R117 must be adjusted several times. Final adjustment of R116 and R117 is as follows.

6. Connect comb generator.

7. Turn R108 and check that the 80-, 90-, and 100-MHz signals move to the left. Readjust R117 so only these signals move to the left.

8. Turn R108 to maximum resistance and adjust R114. All signals 20 to 100 MHz should move to the left. Adjust R116 so this condition can be met.

9. Alternately adjust R116 and R117 so that amplitude corrections are made at the proper frequencies.

10. Turn R106 through R115 to maximum resistance.

11. Determine the lowest frequency needing correction and adjust the appropriately labeled pot. Always start adjustment at the lowest frequency because all frequencies above are affected.

**Fig.** 18 shows final alignment. The camera lens did not have a wide enough angle to display 100 MHz. Note that 10 through 40 MHz are slightly high and 60 through

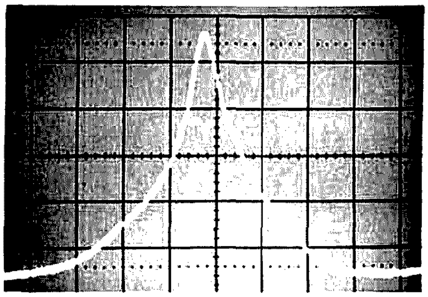


fig. 18. 10-kHz bandwidth position after alignment. Horizontal scale: 10 kHz per division.

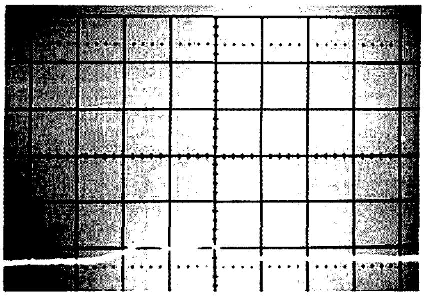


fig. 19. Setup as shown in fig. 24. Turn on video filter and note reduction of noise.

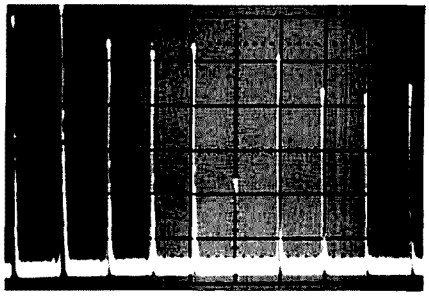


fig. 20. Output from comb generator, 10-100 MHz. Scope camera did not have enough width to display 100 MHz. Maximum errors are about 2 MHz.

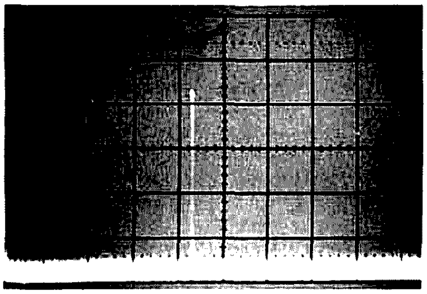


fig. 21. Setup is as shown in fig. 27. I-f gain has been reduced to take advantage of the full dynamic range.

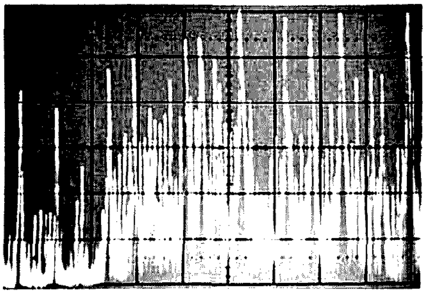


fig. 22. After connecting antenna to inputs, turn BW to 1 kHz and note how stations in the broadcast band can be resolved.

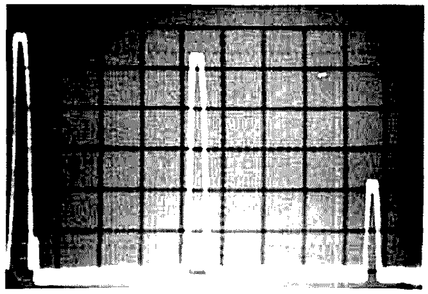


fig. 23. Second harmonic of transmitter operating at about 4 MHz. Second harmonic is down only about 32 dB.

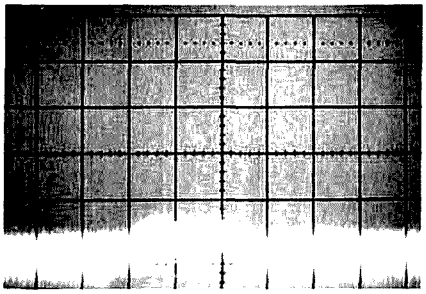


fig. 24. BW 250 kHz, Scan 100 kHz/division and generator frequency about 40 MHz. Increase generator output to see a signal in the noise with the video filter off.

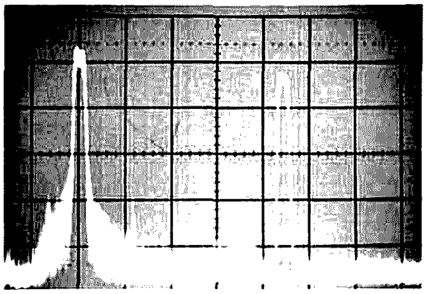


fig. 25. Setup as shown in fig. 27. Set scan to 1 MHz/division. Note video carrier, color information about 3.6 MHz above video carrier, and sound carrier 4.5 MHz above video carrier.

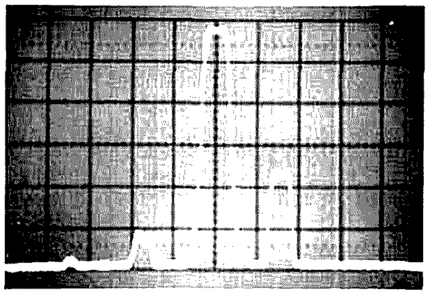


fig. 26. Output of a commercially made ssb transmitter into a 50-ohm load with carrier inserted at about 3.8 MHz. Scan 500 kHz/division. Note spurious signals at about 3050 kHz and 4550 kHz.

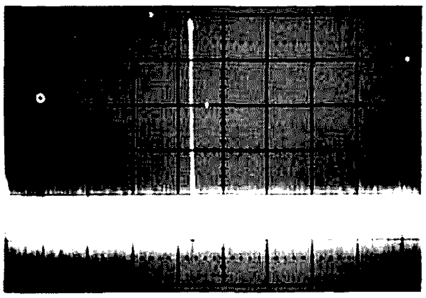


fig. 27. I-f gain adjustment. BW 250 kHz, Scan 10 MHz/division, Generator frequency 40 MHz; generator output about 4 divisions. The i-f gain is set too high, which shifts the baseline up two divisions. This results in a loss of about 20 dB of dynamic range.

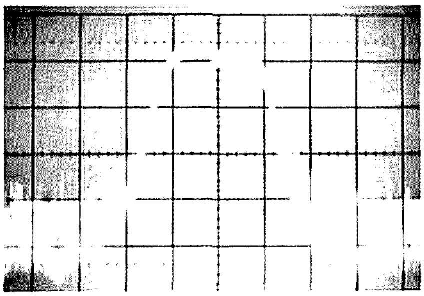


fig. 28. TV station. BW 250 kHz, Scan 100 kHz/division. Adjust frequency control to the video carrier of any TV station between channel 2 and channel 6. Connect 50- or 75-ohm TV antenna to input. Line lock the scope and note video sidebands and vertical sync pulse, which will drift through the top of the signal.

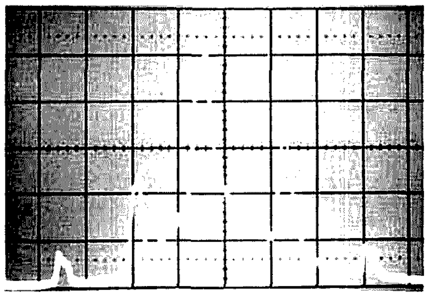


fig. 29. Nonlinear modulation. BW 1 kHz, Scan 10 kHz/division, generator frequency about 40 MHz, and about 50% modulation with a 15 kHz signal. The generator used here was not designed to be modulated at a 15-kHz rate. Note unsymmetrical sidebands and harmonics of the modulated frequency. The harmonics were generated in the modulation process.

90 MHz are slightly low. Maximum error shown here is about 2 MHz.

**Crystal-filter alignment**. Set controls and test equipment as follows:

Bandwidth: 1 kHz.

Scan: 50 kHz/division

Generator frequency: 20 MHz (not critical).

Generator output: 5½ divisions on display.

Sweep speed: 20 ms/divisions.

Video filter: 10 kHz.

Video gain: maximum.

The crystal filters are aligned one at a time. The filter not being aligned should have a 0.01-$\mu$F disc capacitor soldered across the crystal. Always maintain about 5½ divisions on the display but never more than six divisions.

**1.** Adjust C402 for a narrow peak at the crystal-filter frequency.

**2.** Go between 1 kHz and 10 kHz bandwidth, adjusting C401 for narrowest bandwidth without regard to amplitude. Do not adjust C402 in the 10-kHz position.

**3.** Adjust the second crystal filter as you did the first one.

**4.** Increase input level to 10,000 $\mu$V.

**5.** Turn i-f gain down to 5½ divisions and remove both .01-$\mu$F capacitors.

**6.** Alternately adjust C401 and C403 for minimum width and symmetrical skirts at the base of the signal in both the 1-kHz and 10-kHz positions. Possibly there will be some spurious resonances slightly above the filter frequency after alignment in the 10 kHz position.

**Gain compensation.** Set controls and test equipment as follows:

Bandwidth: 250 kHz.

Scan: 50 kHz/division.

Generator frequency ≈ 20 MHz.

Generator output: 5 divisions on display.

Sweep speed: 50 ms/division.

**1.** Go to 10 kHz bandwidth and adjust R403 for 5 divisions.

**2.** Go to 1 kHz bandwidth and adjust R402 for 5 divisions.

**3.** Drill access hole for L401 in i-f amplifier cover and install cover.

**Tuned-circuit alignment.** All tuned circuits are peaked during this procedure. Even though most have been adjusted, this is a good check to see if all are peaked at full operation. Again, only slight adjustments of the 50- and 200-MHz bandpass filters should be needed. If major adjustment is required, or if one coil won't tune, go back to the procedure in the rf section that describes initial alignment. Set controls and test equipment as follows:

Bandwidth: 250 kHz.

Scan: ≈ 500 kHz/division

Generator frequency: ≈ 20 MHz.

Generator output: ≈ 4 divisions on display.

**1.** Peak C301, C302, and C303 for maximum signal.

**2.** Peak C304 through C308 for maximum signal consistent with a flat top across signal.

**3.** Peak L401 for maximum signal consistent with a flat top across signal. There may be about a 1-dB dip in the center of the signal.

**Log amplifier.** Set controls and test equipment as follows:

Bandwidth: 10 kHz.

Scan width: 10 kHz/division.

Generator freuqency: ≈ 20 MHz.

Sweep speed: 10 ms/division.

Video filter: 10 kHz.

**1.** Set the generator output to 30 $\mu$V.

**2.** Set i-f gain for one division. Note that each pot adjusts the level for two divisions and the best compromise should be achieved. The tolerance is ±0.2 division. Always start from one division when performing this alignment.

| divisions | output from generator (microvolts) | alignment pot |
|---|---|---|
| 1 | 30 | R503 |
| 2 | 100 | R503 |
| 3 | 300 | R502 |
| 4 | 1000 | R502 |
| 5 | 3000 | R501 |
| 6 | 10,000 | R501 |

## operation

This section provides an explanation of the function of the controls and concludes with some experiments to demonstrate operation.

**Rf attenuator.** The purpose of this circuit is to reduce the amplitude of the incoming signal to a convenient level for display on the spectrum analyzer. The maximum level to the attenuator should be no greater than 1 volt rms, and the maximum level to the first mixer should be no greater than 10 mV rms. One of the most common errors in the operation of a spectrum analyzer is to use too much i-f attenuation and feed an excessively high level to its input. This results in overloading the rf section, which generates spurious signals and causes possible damage to the first mixer. Never overload the front end.

**Frequency control and fine tuning.** The frequency control tunes the spectrum analyzer to the desired operating frequency. In the 10 MHz/division position, the center frequency is set to 50 MHz and the frequency control is inoperative. The fine tuning control has a range of about 500 kHz.

**Scan width.** After setting the center frequency with the frequency control, the scan width determines the frequency width that will be displayed. For example, if the frequency control is set to 20 MHz and the scan width to 100 kHz/division, the spectrum analyzer will sweep from 19.5 to 20.5 MHz.

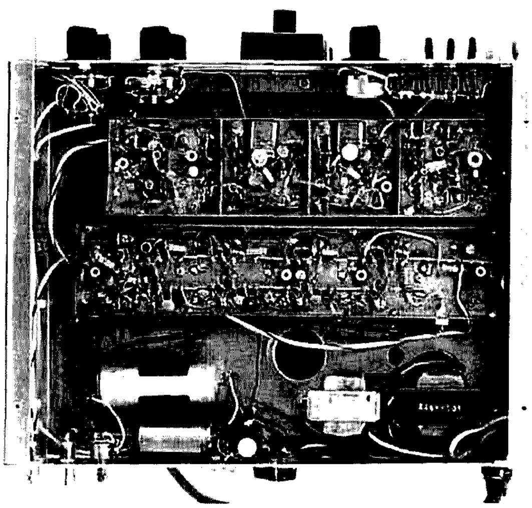
**Bandwidth.** As in a conventional radio receiver, this control determines the selectivity of the analyzer. Narrow bandwidth is required to separate signals relatively close together and wide bandwidth is required for high sweep speeds.

**I-f attenuator and i-f gain.** The i-f attenuator is used for limited tests where discrete steps of attenuation are required. The i-f gain is used to provide sufficient gain with minimum noise at the baseline. The noise at the baseline rises as the bandwidth is increased and should be maintained at between one-half and one division to avoid degrading the dynamic range. See figs. 22 and 23.

**Video filter.** The video filter is used to remove high-frequency noise from the signal when looking at small scan widths with slow sweep speeds. See table 2 for video filter operation. Figs. 24 and 25 are examples of its effect in the 250-kHz bandwidth position.

**Resolution.** If the sweep speed is too high, the scan (or sweep) width too great, or the bandwidth too narrow, the display will lose amplitude or smear. Figs. 26 and 27 show displays of a broadcast signal at the 250-kHz bandwidth setting. Further examples are shown in figs. 28 and 29, which are displays of a local television station.

A quick check on resolution is to decrease sweep speed and look for narrowing or increased amplitude of the displayed signal. The sweep speed should never exceed 2 ms/division. A P2 and P7 scope tube would be advantageous but is not absolutely required. Table 2



Bottom view of spectrum analyzer chassis showing the i-f assembly (along front of chassis, top) and logarithmic amplifier (center). Power supply components are located along the rear of the chassis, bottom.

table 2. Maximum sweep speeds for resolution at spectrum-analyzer bandwidth settings.

| bandwidth (kHz) | scan width (MHz/division) | maximum sweep width (per division) | video filter (kHz) |
|---|---|---|---|
| 250 | 10 | 10 ms | Off |
| 250 | 1 | 5 ms | 10 |
| 250 | 0.1 | 2 ms | 10 |
| 250 | 0.01 | not usable | — |
| 10 | 10 | not usable | — |
| 10 | 1 | 0.2 sec | 10 |
| 10 | 0.1 | 50 ms | 10 |
| 10 | 0.01 | 20 ms | 1 |
| 1 | 10 | not usable | — |
| 1 | 1 | not usable | — |
| 1 | 0.1 | 0.2 sec | 10 |
| 1 | 0.01 | 50 ms | 1 |

gives maximum sweep speeds for some scan widths, bandwidths, and settings of the video filter.

**400-500 MHz operation.** If the input lowpass filter is disconnected, the 200- to 300-MHz first local oscillator will mix with 500 to 400 MHz, producing the required 200 MHz first i-f. This was tried and some strong local low-frequency signals leaked through. Sensitivity seemed similar to normal operation, but signals were displayed in reverse.

**Use as a radio receiver.** The output can be connected to an audio amplifier with 1-megohm input impedance and used to monitor radio signals. However, because it is a spectrum analyzer, the instrument has several limitations when used as a radio receiver. Dial calibration is accurate to only ±3 MHz. Because of its high frequency, the local oscillator drifts and some fine tuning might be required. There is no tuned circuit at the front end, therefore a strong signal can cross-modulate the analyzer. Despite these shortcomings, WWV, BBC, Australian Broadcasting, Radio Netherlands, Radio Moscow and many other stations have been received using a TV feedline as an antenna.

As with most anything else, there is more than one way to design a spectrum analyzer. If you find an easier or better way to improve this spectrum analyzer, without degrading its performance, or find any obvious errors, I'd appreciate hearing from you.

## reference


1. Wayne C. Ryder, W6URH, "500-MHz Decode Prescaler," *ham radio*, June, 1975, page 32.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1. *Spectrum Analyzer Basics*, Hewlett-Packard Applications note 150, Hewlett-Packard, 1501 Page Mill Road, Palo Alto, California 94304.
2. *Spectrum Analyzer Measurements*, (Tektronix part no. 062-1070-00), Tektronix, Inc., P.O. Box 500, Beaverton, Oregon 97077.
3. *Spectrum Analyzer Circuits*, (Tektronix part no. 062-1055-00), Tektronix, Inc., P.O. Box 500, Beaverton, Oregon 97077.


**ham radio**

# RTTY tape editor

This tape editor is designed for use with a transmitter-distributor (or TD as it is more commonly known), a keyboard send-receive teletype (KSR) and a reperforator. Any serious RTTY operator who owns this equipment has surely longed for an easy way to delete or insert material into a tape. This need could arise from errors in reception when punching tape off the air, or it might be merely a desire to modify a tape to suit your particular needs. In any case the arrangement described here should fit the bill.

## tape editing

The usual method of tape editing is to insert the original tape into the TD and, while operating on local loop, make a new tape from the original tape on the reperforator. Editing is done by stopping the tape in the TD, entering corrections from the keyboard, moving the tape manually through the TD past the incorrect material, then restarting the TD to continue until reaching the next point for correction. The result of this exercise is monitored on the page printer. The problem with this procedure is getting the original tape moved to just the exact spot from which it should be restarted. This may sound simple enough, but if you've ever tried this method you know how easy it is to end a character or so off from where you intended to restart, with another error to edit.

## examples

With the tape-edit circuit described here, the TD is stepped one character for every character entered from the keyboard. Thus, if the tape reads "shop," and you wish to change it to "ship," you can stop the TD after "sh," put the *EDIT* switch to *ON*, and type "i." The TD will advance one character when the "i" is entered so that when the *EDIT* switch is thrown to *OFF*, the TD will start again and print the letter "p" thus completing the correction. Alternatively, the tape could have been stopped just before the word "shop," the *EDIT* switch placed to *ON* and the entire word "ship" typed in from the keyboard. The TD would have been stepped along one character for each character typed so that the net result would be the same.

Suppose the word "ship" had been garbled in the original tape, but from the message context you were able to determine that "ship" was the correct word. Suppose it read "shxmp." Stop the tape as before and turn the *EDIT* switch to *ON*. Enter "i" from the keyboard then strike the *LETTERS* key, which will move the TD one more character past the extraneous "m." Turn the *EDIT* switch to *OFF*, which will restart the TD with the letter "p," and the correction will be complete. With these corrections, editing can be done on the air after a very little practice.

The circuit is shown in **fig. 1**. The objective is to advance the TD one character for each character entered from the keyboard without having the TD affect the print (or the new tape being punched by the reperforator).

A miniature dpdt toggle switch was mounted on the front of the TD just below the main *ON/OFF* switch and slightly to the left. This required drilling a ¼-inch (6.5mm) hole in the TD casting – an easy job. One section of this switch, S1A, was placed in parallel with the TD signal line so that a *MARK* condition exists at all times when the *EDIT* switch is in the *ON* position. The other section, S1B, is connected in series with the *TAPE-OUT* and the *TAUT-TAPE* switch and is closed when the *EDIT* switch is *OFF*. A pair of leads parallel S1B and run to the contacts on a polar relay. When the *EDIT* switch is *ON*, S1B is open so the only way that current can get to the *RUN MAGNET* on the TD is through the polar-relay contacts.

The polar relay coil is connected in series with the keyboard. Thus loop current flows through the coil until a key is depressed. This action holds the relay in the *MARK* position.

When a key is depressed, the *START* pulse moves the polar relay to the *SPACE* position, which closes the circuit to the TD run magnet, allowing the TD to start. As soon as the *STOP* pulse is transmitted for whatever character was typed in, the polar relay will return to the *MARK* position, which opens the run-magnet circuit, stopping the TD after it has advanced one character. Thus the TD advances the tape one character for each character sent from the keyboard. Since the first section of the *EDIT* switch has shorted the TD signal line, the TD has no effect on the transmission.

## tape changes

Earlier in the article I discussed error corrections. Sometimes it's desired to make changes in a tape. For example I have a Christmas-greeting tape, which was received from the State Council of Civil Defense. I wanted to use this tape with my name in the box where the original read "State Council of Civil Defense." **Fig. 2** shows the print before and after editing.

## cautionary measures

Remember that nonprinting functions must be taken into account when editing. For example, if W3EAG is to be replaced by "Tommy," this is not just a one-for-one substitution as it appears. The figure function preceding the 3, and the letters function following it, must be accounted for since these are two additional characters

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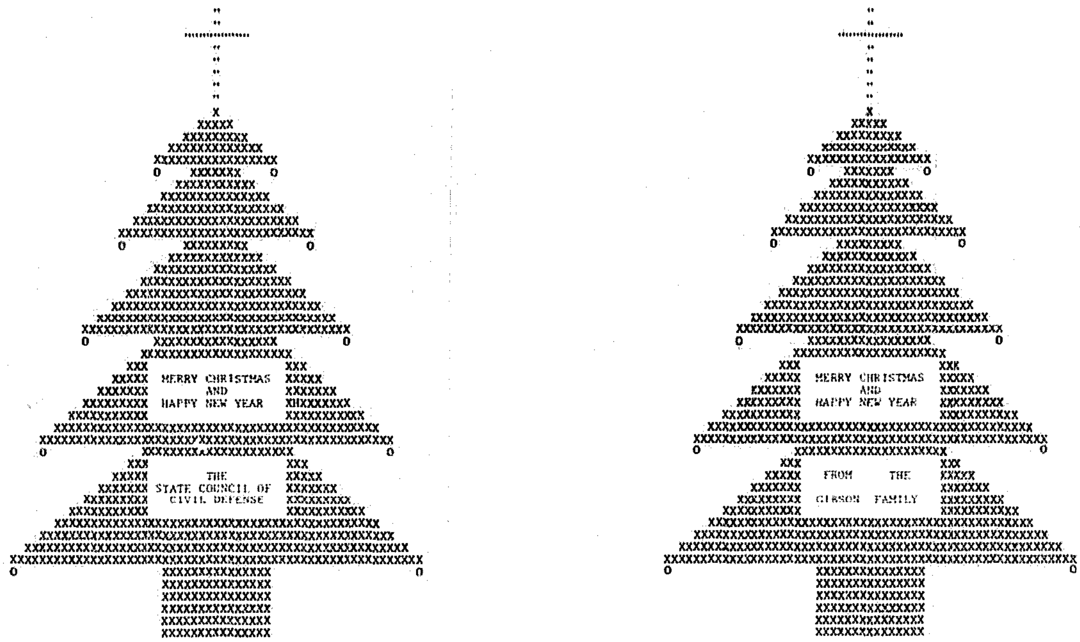


**fig. 2. A Christmas message before and after editing.**

on the original tape that are not required for this particular substitution. This can be accommodated by striking the *LETTERS* key twice, which will advance the tape but not the printer. However, if W3EAG were being replaced with K2XXX this would be a one-for-one substitution and no *LETTERS* need be struck.

Always stop the TD with the *TAUT-TAPE* switch, then turn the *EDIT* switch to *ON*. If you attempt to stop the TD by turning the *EDIT* switch to *ON*, you may garble a character since the edit switch shorts the TD signal line.

The polar relay used in my rig switches ac to the run magnet. Some operators may be using dc, in which case some attention to spark suppression at the relay contacts would be advisable.

**ham radio**

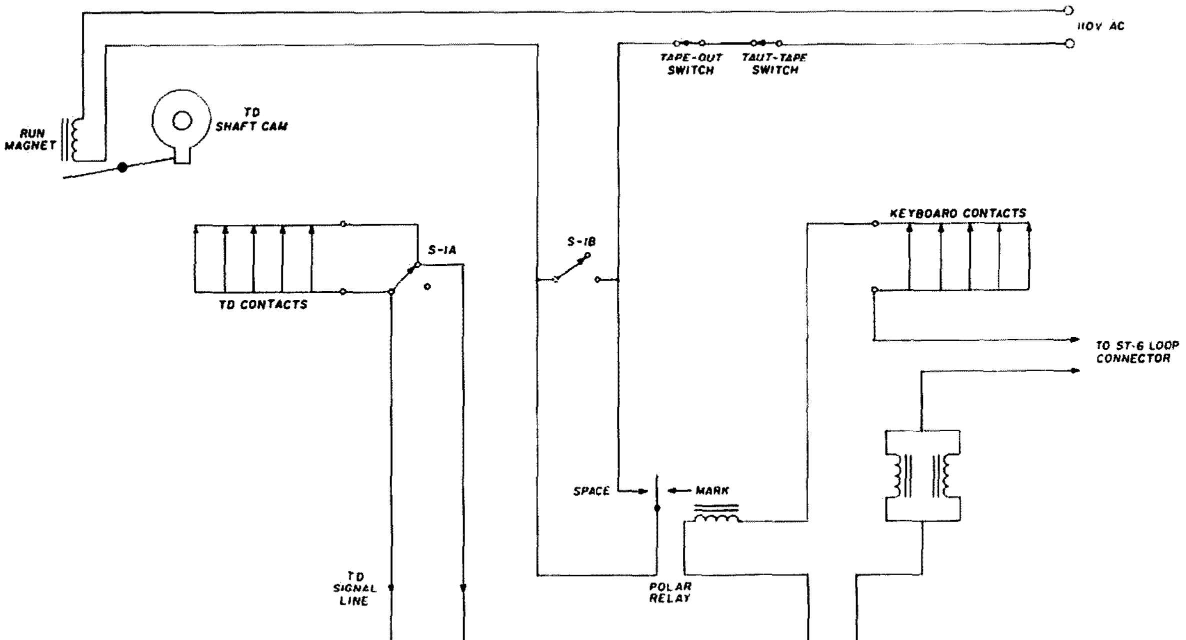


**fig. 1. Tape-editing circuit, which allows error corrections and tape changes.**

# top-coupled bandpass filter

## a chebyshev design

Practical design and construction data for 3-, 4- and 5-section Chebyshev bandpass filters for amateur applications

The **top-coupled bandpass** filter is the easiest of its type to construct. It appears to be several parallel resonant circuits connected in sequence with small coupling capacitors. All inductors are the same value. In actuality, this type of bandpass filter is one of many configurations based on a lowpass filter prototype. Presented here are the necessary component formulas to enable design and construction of 3-, 4-, and 5-section "1-dB ripple" Chebyshev bandpass filters along with passband and skirt selectivity response, input impedance, and insertion loss.

### why Chebyshev?

The most common filter design is the Butterworth with its flat passband response. A Chebyshev filter has ripple in the passband, defined by dB difference of peak-to-valley; design bandwidth refers to the dB-down points on the edges rather than the 3 dB points of the Butterworth. The Chebyshev response also has better skirt attenuation than the Butterworth.

The names refer to the mathematical operations used in determining component values. The math involved is lengthy and may be found in many textbooks. Here, the only math will be that required for one type of Chebyshev ripple design.

Both Chebyshev and Butterworth designs for equal load and source impedances suffer shape distortion and increased insertion loss as a function of finite Q. This will be seen on the response curves.

Schematics and individual value formulas are shown in fig. 1. The formulas are derived from lossless-element designs and feature equal end resistances. Following are definitions common to all:

$f_L$ = *Lowest frequency of passband, MHz*
$f_H$ = *Highest frequency of passband, MHz*
$f_o$ = *Geometric center frequency* = $\sqrt{f_L f_H}$
$f_b$ = *Bandwidth* = $f_H - f_L$
$p$ = *Fractional bandwidth* = $f_b/f_o$
$C_o$ = *Resonating capacitance* = $\dfrac{25330.3}{f_o^2 L}$

with $L$ equal to $L3$, $L4$, or $L5$, depending on filter

Note that all terms are scaled to megahertz, micro-henries, picofarads, and kilohms. Practical designs require high end resistances. Matching to lower values is given later.

### how degrading can you get?

Circuit Q has little effect on skirt attenuation but does distort the passband. Skirt response is shown in **figs. 2** and **3** while passband response is shown in **figs. 4, 5**, and **6**.

The parallel $L$-$C$ combined Q will determine the passband shape and insertion loss. It can be shown that

$$Q = \textit{combined } Q_C \textit{ and } Q_L = \frac{Q_L Q_C}{Q_L + Q_C}$$

A reference to Q will generally refer to the combination. Very high Q will still result in distortion, as shown on the passband curves. The curves are *not* ideal — they have been deliberately chosen to represent degraded performance in choosing coils and coupling capacitors as will be explained in the design example.

This unorthodox presentation is the "worst-case" design situation. Selection of component values closer to calculated values will improve shape, insertion-loss, and skirt attenuation. Curves of **figs. 2** through **6** were taken from the design example where the inductor was higher in value than given by the formula, the coupling

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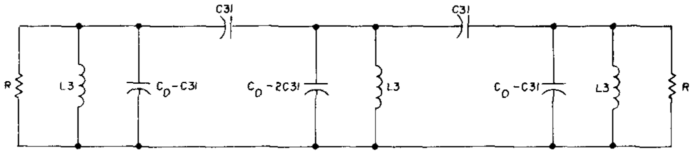
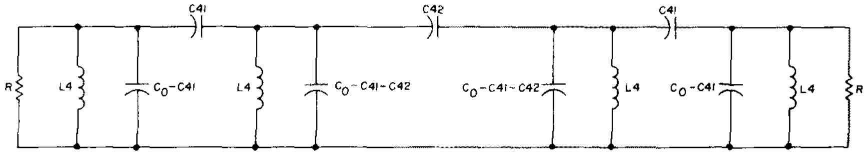


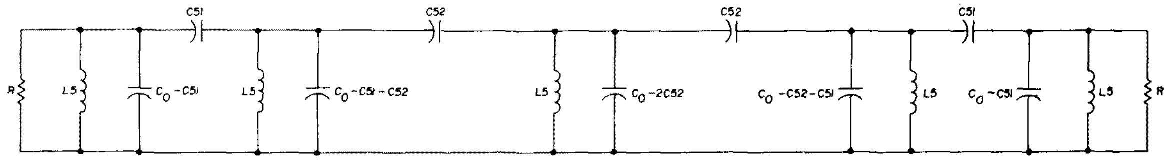

$$L3 = \frac{786\ p\ R}{F_o} \qquad C31 = \frac{227}{F_o\ R}$$

3 SECTION




$$L4 = \frac{765\ p\ R}{F_o} \qquad C41 = \frac{225}{F_o\ R} \qquad C42 = \frac{206}{F_o\ R}$$

4 SECTION




$$L5 = \frac{745\ p\ R}{F_o} \qquad C51 = \frac{223}{F_o\ R} \qquad C52 = \frac{188}{F_o\ R}$$

5 SECTION

**fig. 1. Basic top-coupled bandpass filters of Chebyshev design. All values are in MHz, microhenries, pF and kilohms. The value p is the fractional bandwidth, as discussed in the text.**

### general philosophy of design

The intention of the design data presented here is to enable the nonspecialist to obtain working values as quickly as possible in a wide area of applications. The "symmetric" design of equal source and load resistance meets this criterion even though the component calculations are for lossless elements. The trade-off of increased insertion loss and passband distortion was considered valid for a non-commercial application when compared with ease of calculation and impedance matching.

The filter specialist may prefer some more "meat," but this requires reference to heartier texts and references. It would have been nice to present a table of filter elements – this is possible by using Craig's excellent *Design of Lossy Filters* handbook (see bibliography). It was considered for this article, then shelved for several reasons. First, there is the choice of bands; which frequencies to choose besides the obvious 160- to 6-meter bands, such as for intermediate frequencies and the like. Secondly, what would be the expected impedance levels? The real world is not always 50 ohms. Last, such designs require a specific unloaded Q of resonators which can be achieved by resistor loading of higher-Q components; the passband takes on strange shapes when this is not done.

An optimum shape and loss design takes at least twice the effort to obtain working hardware, even with a table of values to start with. The prime consideration was the amount of time available to the Amateur – calculations can be done during your spare time, lunch breaks, etc., but working with hardware requires workshop time.

capacitor lower. This is the general rule in component selection. The Butterworth 3-section passband of **fig. 7** is based on *ideal* values and is useful for comparison.

Decreasing Q will narrow the bandwidth. This does not mean that a low Q is unusable – simply that the design bandwidth must be changed if it does not fit a particular situation. This can be used to advantage since *required* unloaded Q is inversely proportional to bandwidth.

## scaling

The horizontal scale of **figs. 2** through **7** may be read in two ways: Units of bandwidth plus or minus center frequency as a universal value, *or* fractional frequency as a function of center frequency for a 5 per cent bandwidth. The curves may be used with bandwidths from 3 to 8 per cent of center frequency with reasonable accuracy.

The inverse relationship of Q versus bandwidth can be seen by examining required inductance value for a given load resistance. Inductance must increase with bandwidth because of the $p$ term in the numerator of the formula. Since the loss from a finite Q appears as a resistance in parallel with the shunt $L$ and $C$ at a value of Q times the reactance of either at center frequency,

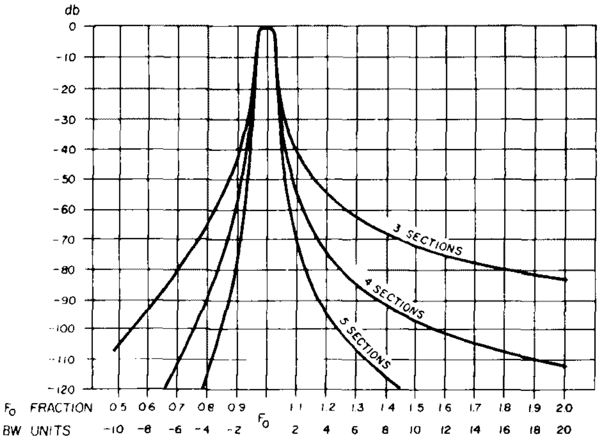



**fig. 2. Skirt response of 1-dB ripple Chebyshev filters with 5% bandwidth, worst case example.**

increasing the reactance will also increase the equivalent resistance. It is the ratio of equivalent resistance to the source and load resistances that determines insertion loss and shape change.

If the available $Q$ is 160 and the percentage bandwidth only 2.5, the $Q=80$ curve of the 5 per cent bandwidth example will be the response. Similarly, if the bandwidth is to be 10 per cent, the $Q=320$ curve would be used. Other Qs and bandwidths may be interpolated.

Variations in $Q$ will have very little affect on skirt response below −30 dB; skirt response beyond this can be taken directly from the curves since the reactances and end resistances control here.

## example

A bandwidth of 0.5 MHz with a center frequency of 10 MHz is desired. Source and load resistance will be 5 kilohms. For the 3-section filter,

$$L3 = 78.6 \times 0.05 \times 5/10 = 1965 \mu H$$
$$C31 = 227/(10 \times 5) = 4.54\ pF$$

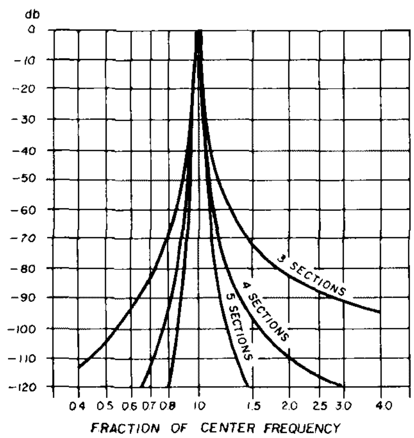



**fig. 3. Skirt response of 1-dB ripple Chebyshev filters with 5% bandwidth, worst case example. This is the same as fig. 2 except that frequency is plotted on logarithmic scale.**

Resonating capacitance $C_o$ is determined solely by L3, L4, or L5 at the center frequency:

$$C_o = 25330.3/(10 \times 10 \times 1.965) = 128.907\ pF$$

The actual parallel circuit capacitance will be $C_o$ minus the adjacent coupling capacitance values, so the end values will be $128.907 - 4.54 = 124.367\ pF$, and the middle value will be $128.907 - 2(4.54) = 119.827\ pF$.

For the 4-section filter:

$L4 = 1.9125\ \mu H$
$C_o = 123.446\ pF$
$C41 = 4.5\ pF$
$C42 = 4.12\ pF$

End parallel capacitors = $127.946\ pF$
Middle parallel capacitors = $123.826\ pF$

For the 5-section filter:

$L5 = 1.8625\ \mu H$
$C_o = 136.002\ pF$
$C51 = 4.46\ pF$
$C52 = 3.76\ pF$

End parallel capacitors = $131.542\ pF$
Next-to-middle capacitors = $127.782\ pF$
Middle parallel capacitor = $128.482\ pF$

All of these designs look very nice on paper but are not practical because none of the component values are commercially available. To obtain practical values, it was decided to select the nearest higher value of $L$ in the 10% tolerance value of $2.2\ \mu H$. Coupling capacitance was selected as the nearest lower 10% value, $3.9\ pF$ for C31, C41, C51, and $3.3\ pF$ for C42 and C52. $C_o$ now becomes $115.138\ pF$ and the actual circuit parallels chosen by the adjacent coupling capacitor subtraction rule.

These values are extreme, particularly for the inductor, but serve to show "worst-case" variations and are the values calculated for the curves of **figs. 2** through **6**.

As a very general statement, lowering coupling capacitance will improve skirt attenuation with little affect on passband ripple; increasing inductance reduces ripple and skirt attenuation. Lowering coupling capacitance and increasing inductance together has the effect of approaching Butterworth response. A simultaneous change in the opposite direction will increase both ripple and skirt attenuation.

A low-Q situation always results in a rounded passband, whether Butterworth or Chebyshev. Skirt attenuation is governed by reactance only at frequencies well removed from center. This can be seen by the converging curves of **figs. 3, 4, 5,** and **6** at 20 *dB* down.

## insertion loss and input impedance

Using lossless elements, the input impedance of all odd-number section filters will be equal to the load resistance at the center frequency. The impedance seen

by a current source (transistor collector or tube plate) feeding the input will be half the load resistance for symmetric load and source resistances.

Input impedance will vary over the passband with magnitudes changing as much as 2:1. Magnitude is maximum where ripple in the passband is at its peak. Odd-section filters have a peak in the middle whereas even-sections have a dip. The four-section filter will have a lower input impedance at the middle but will be equal to load resistance at the peaks. Because of this variation, input impedance must be considered to be a *mean* value over the passband.

Insertion loss, as used here, is the ratio of input voltage magnitude (including the source resistance) to output voltage magnitude as compared to the peak of lossless element response. Insertion loss is commonly the input/output ratio alone but the ripple response of Chebyshev designs requires a different definition.

**Fig. 8** shows the variation of input impedance and insertion loss for ideal values but with finite values of Q. This may be used as a general guide.

## response of the filter

For an interstage application, the driver will have a load equal to source-resistance in parallel with input impedance of the filter. Insertion-loss data assumes use of lossless elements so the driver load will be half the design resistance. For the 5k example, the driver load is 2.5k when computing overall driver and filter gain.

As an example, take the 3-section filter as graphed in **fig. 4** with a Q of 80. Assume the driver stage has a 20 dB gain with a 2.5k load. At the center frequency, filter insertion loss is 5 dB so the overall gain is *20 - 5 = 15 dB*. From **fig. 2**, the skirt attenuation at $0.8f_o$ is *67 dB*, so the overall gain is *20 - 67 = -47 dB*. Relative difference between $f_o$ and $0.8f_o$ is *-62 dB*.

Any driver resistance, such as $1/h_{oe}$ of a transistor, must be part of the load resistance with the actual load adjusted to fit the design resistance. Stray capacitance becomes part of the end section capacitance.

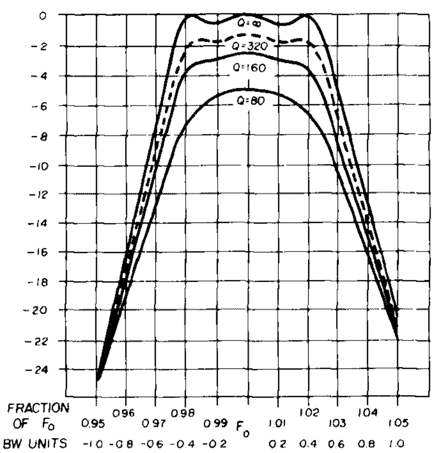



**fig. 4. Passband response of 3-section, 1-dB Chebyshev filter vs Q. Worst case example.**

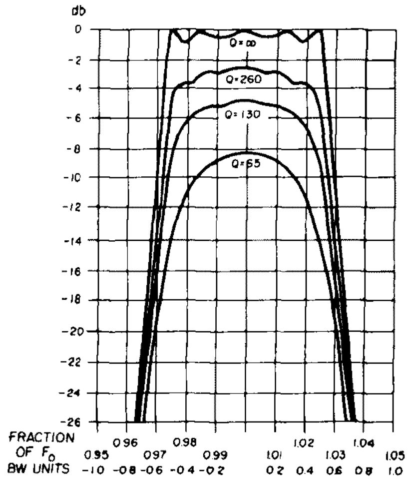



**fig. 5. Passband response of 4-section, 1-dB Chebyshev filter vs Q. Worst case example.**

**Fig. 9** shows three different matching methods. They apply to either end and assume matching to a pure resistance. Filter end resistance varies with section Q and is shown in **fig. 8** as a fractional value of design

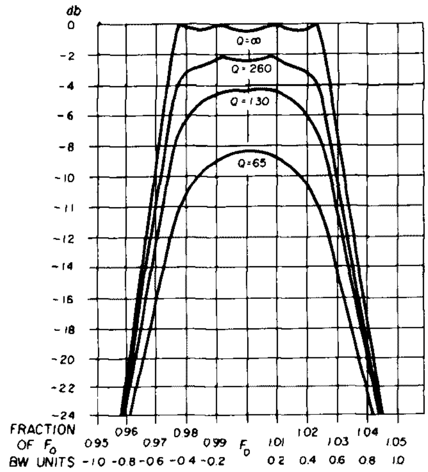



**fig. 6. Passband response of 5-section, 1-dB Chebyshev filter vs Q. Worst case example.**

resistance. The following definitions apply to all methods:

$R_i$ = Actual input resistance; obtain by multiplying design *R* by factor shown in **fig. 8**

$R_g$ = Low resistance to be matched

$X_i$ = Inductor reactance at center frequency less reactance of the adjacent coupling capacitor at the center frequency.

$f_o$ = Center frequency of filter

The inductive-tap method assumes a unity coupling factor and should work if toroids are used with windings

spaced evenly over the entire form. Lower permeability toroids such as Amidon types 2 and 6 should have at least 75 percent of the toroid form filled with wire. Note that wide spacing and placing a gap between start and finish *will lower the coupling factor as well as* $Q$.

Best results in overall performance of capacitive matching is achieved by making $C_s$ and $C_a$ fixed values with $C_p$ and $C_b$ trimmable. The end inductor must still resonate at $f_o$. All of the matching methods will work at either end.

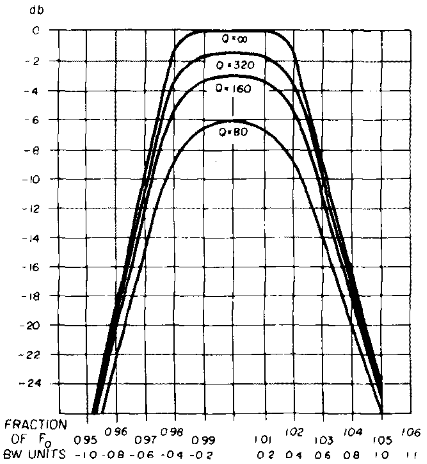



**fig. 7. Passband response of 3-section ideal Butterworth filter vs** $Q$.

The impedance looking into the filter will not be uniform with any of the methods. The inductive-tap and $C_a/C_b$ arrangements will present low impedances outside of the passband; the $C_s/C_p$ type will present high impedances outside of the passband. When used with antenna-inputs, a small tee or pi resistance pad should be used to prevent unusual responses due to mismatch.

**table 1. 50-ohm attenuation pads using standard 10% resistors.** $R_{in}$ **assumes 50-ohm load; worst vswr is with output open or shorted.**

| | TEE-pads | | | |
|---|---|---|---|---|
| dB | R1 | R2 | $R_{in}$ | worst vswr |
| -3.52 | 10 | 120 | 50.00 | 2.60 |
| -4.24 | 12 | 100 | 50.27 | 2.24 |
| -5.29 | 15 | 82 | 51.26 | 1.94 |
| -6.68 | 18 | 56 | 48.71 | 1.58 |
| -8.14 | 22 | 47 | 50.44 | 1.38 |
| -10.47 | 27 | 33 | 50.10 | 1.20 |
| -13.64 | 33 | 22 | 50.39 | 1.10 |
| -18.43 | 39 | 12 | 49.57 | 1.04 |
| | **PI-pads** | | | |
| dB | R3 | R4 | $R_{in}$ | worst vswr |
| -1.74 | 470 | 10 | 49.39 | 5.11 |
| -2.05 | 470 | 12 | 50.99 | 4.76 |
| -2.58 | 330 | 15 | 49.63 | 3.48 |
| -3.75 | 220 | 22 | 48.82 | 2.50 |
| -4.56 | 180 | 27 | 48.36 | 2.13 |
| -6.19 | 150 | 39 | 50.66 | 1.67 |
| -9.66 | 100 | 68 | 50.33 | 1.25 |
| -12.51 | 82 | 100 | 50.44 | 1.13 |
| -15.85 | 68 | 150 | 49.27 | 1.07 |

A table of attenuation pads using 10-percent tolerance resistors is shown in table 1.

Voltage gain is achieved when matching is only at the input; voltage loss occurs when matching is on the output. This is true for all three types. Gain or loss is equal to the square-root of $R_i/R_g$. Filter loss is still present and the curves of fig. 8 should be used for the overall gain or loss.

To illustrate this, assume the 3-section example already given with a $Q = 160$. From fig. 8, the loss of the filter alone is 2.6 dB and $R_i = 3.74k\ (0.718 \times 5k)$. The input $R_g$ is $50$ *ohms* so $R_i/R_g = 7.48$ with a voltage loss of $1/2.735$ or $-8.74$ *dB*. The net voltage gain is $18.74 - 2.6 - 8.74 = +7.4$ *dB*.

## capacitive matching examples

Before beginning, the following reactance formulas apply with all values in ohms, pF, μH, and MHz:

$$X_L = 6.28319 f_o L$$

$$X_C = \frac{159155}{f_o C}$$

Take the 3-section example with $Q = 160$ and match to 50 ohms. $R_i$ is found to be 3.74k from the fig. 8 curve and $R_g$ will be 50 ohms. $L_i$ is *not* the inductor value used but is modified by the coupling capacitance immediately adjacent, plus any distributed capacitance of the inductor. The latter can be discounted since a 2.2 μH inductor will have little distributed capacitance. A properly-adjusted filter will cancel all reactance at the center frequency since each coil will resonate with the combination of parallel capacitance and adjacent coupling capacitance. All of the matching component formulas take this into account.

Consider the $C_a/C_b$ arrangement, sometimes called the "capacitive voltage divider" match. To find $X_i$, the reactance of the inductor, $X_L$, must be found (138.230 ohms at 10 MHz for 2.2 μH); $X_{Cc}$ represents the coupling capacitance of 3.9 pF which will be 4080.90 ohms. Therefore,

$$X_i = \frac{X_{Cc} X_L}{X_{Cc} - X_L} = \frac{4080.9\,(138.23)}{4080.9 - 138.23}$$

$$= 143.08\ ohms$$

For this type of match, intermediate terms are calculated to simplify computation:

$$b = \frac{R_i X_i}{R_i^2 + X_i^2} = \frac{3740\,(143.08)}{(3740)^2 + (143.08)^2} = 0.0382$$

$$d = bX_i = 0.0382\,(143.08) = 5.465$$

Then:

$$X_{Ca} = R_g \sqrt{\frac{d}{R_g - d}} = 50 \sqrt{\frac{5.465}{50 - 5.465}} = 17.516\ ohms$$

$$X_{Cb} = bR_i - \sqrt{d(R_g - d)}$$

$$= 0.0382(3740) - 5.465(50 - 5.465)$$

$$= 142.867 - 243.403 = 127.266\ ohms$$

From the reactance formulas, $C_a = 908.6\ pF$ and $C_b = 127.3\ pF$. The nearest 5% tolerance value for $C_a$ is 910 pF but any combination within 5 percent is adequate.

The $C_b$ value is from the exact solution and would be a trimmable value in actual construction. $C_b$ could also be found from

$$C_b = \frac{C_r C_a}{C_a - C_r}$$

where $C_r$ is the required resonating value. In this case $C_r = 115.1 - 3.9 = 111.2\,pF$. If $C_a$ was fixed at 910 pF, $C_b = 126.7\,pF$. The $C_b$ formula above is approximate but is very close to actual.

The $C_s/C_p$ matching arrangement is simpler to calculate. For the same example

$$X_{Cs} = \sqrt{R_g(R_i - R_g)} = \sqrt{50(3740 - 50)} = 429.5\ ohms$$

From the reactance formula $C_s = 37.1\,pF$ but a fixed 39 pF value will work well. The simple computation of $C_p$ results in values very close to actual and is given by

$$C_p = C_r - C_s = 111.2 - 39 = 72.2\,pF$$

using the fixed value of $C_s$ as 39 pF. As in the other case, $C_p$ would be trimmable. If the $C_a/C_b$ matching system is used in interstages, supply voltage bypass capacitors should be at least a hundred times larger than $C_a$.

## construction

Each section of the filter, the parallel component group, should be shielded from every other section. Even if toroids are used the shielding must be used since they can couple slightly by the magnetic field, more so by electrostatic coupling. Only the coupling capacitors should be common to adjacent sections.

Double foil, one-ounce printed-circuit board material is suitable if all the joints are completely soldered. One large piece can be used as a baseplate with strips forming the sides and interstage shields. Threaded spacer rods can

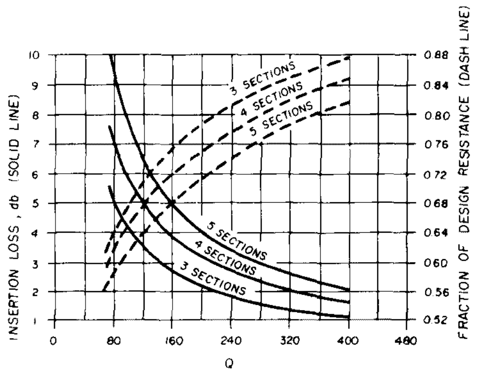



**fig. 8. Ideal value insertion loss and end resistance for 3-, 4-, and 5-section, 1-dB Chebyshev bandpass filters.**

be used in the corners of section compartments, soldered to the board material. With this construction, section tops can be made removable; copper tape should be used on removable tops to insure 100 dB shielding.

## tuning methods

Fig. 10 shows a setup for tuning each filter section. This may be done as the filter is built, section by section. The loose coupling for end sections must be

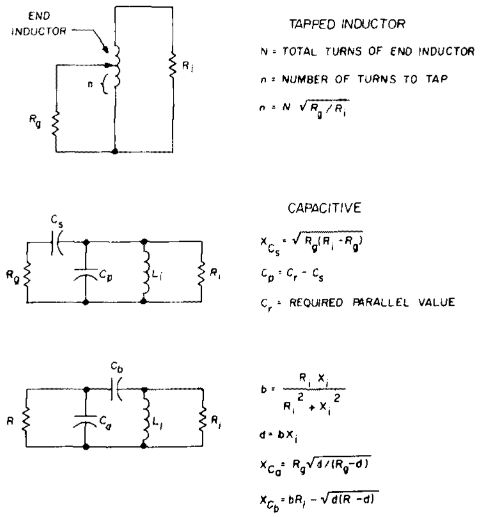



**fig. 9. Impedance matching methods for bandpass filters. Complete information on calculating component values is given in the text.**

*very* loose, particularly at high frequencies, to avoid stray capacitance that can cause mistuning.

An easier method can be used by the fact of passband rounding at low values of Q. This is simply loading of each section by a 1/4-watt carbon composition resistor of about the same value as the design resistance. All the trimmers are then peaked at the center frequency. Several passes should be made since there is some interaction until all capacitances are balanced.

Fig. **10C** shows the setup for coupling the high impedance ends (500 ohms or higher) with the simple tuning method. This is the "Norton Theorem" transformation from a current source to a voltage source with the 10- and 39-ohm resistors serving to isolate the generator and receiver impedances. For 50-ohm end matches, connect directly but use at least 10 dB of attenuation at each end for isolation.

When the simpler method of tuning is completed, the loading resistors are removed and replaced by megohm-range carbon-composition resistors of the same wattage and brand. The reason for replacement is that most all carbon resistors have about 1 or 2 pF of shunt capacitance but the variation between different resistance values of the same wattage and brand is very small. This prevents mistuning by the added capacitance. Film-type resistors have been found to vary greatly with frequency, and some even exhibit inductance instead of capacitance. Where the parallel circuit capacitance is about 400 pF or greater, replacement is not needed since the tuning change is very small.

## component Q

Manufacturer's data on toroid Q is quite reliable. In many instances the required inductance needs fewer turns than table data. To keep Q at its highest value, always use the entire form with even spacing and a minimum gap between the winding start and finish. Use the largest wire size that will fit the form.

Coating of the finished coil is recommended and

ordinary exterior varnish is very good. A double light coating will reduce Q by 7 to 8 per cent. Polyurethane varnish can result in a Q reduction of about 12 per cent and is not recommended. Acrylics and *Q-dope* are generally useless because, in time, moisture can cause lifting of the adhesion area. Acrylics look pretty and are fast-drying but need porous or laquer-compatible surfaces for long life. Q reduction is about the same as varnish.

Dipped-mica and mica-compression trimmer capaci-

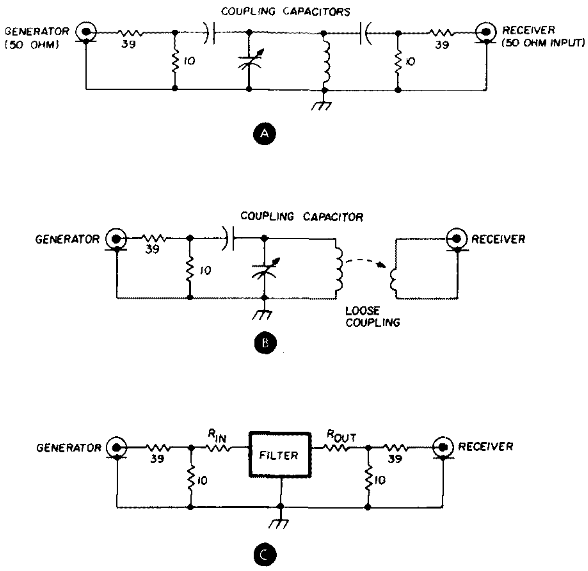



**fig. 10. Test setups for tuning the bandpass filters for the desired passband response.**

tors have *Qs* of 1000 or more in the 3- to 30-MHz range. Silver micas are good at low frequencies but a few have been found to have a Q as low as 600 at 30 MHz. However, using a Q value of 1000 for dipped-mica types is a good rule. Paralleling mica capacitors will not generally affect Q of the total.

If the capacitor is above about 1000 pF, Q may be lower and may also vary with different types. It is best to measure the higher capacitance values.

Shielding will affect inductor Q. Data on shielding affects of single-layer solenoid coils can be found in many texts. Toroids have been found to exhibit negligible Q reduction if kept at least one form-thickness away from the shielding on any surface. Foam polystyrene used as packing material is a good support in this case.

## test setups for Q measurements

Fig. 10A can be used for parallel resonance and Q tests if the coupling capacitors are replaced with 1 pF units above 3 MHz, about 10 pF below. A 1-pF capacitor can be made from a 1/4-inch (6.5mm) square of double-side G10 fiberglass circuit board material. In fact, small-value coupling capacitors can be made this way for higher-frequency filters as part of the interstage shielding but the thickness should be measured and dielectric constant known; keep edges at least 1/8th inch (3mm) from the ground surface and make certain the grounded area of the shield completely surrounds it.

A requirement of this setup is the ability of the receiver to measure 3 dB differences in amplitude from the peak. The signal generator must have a counter attached for accurate frequency differences. If the generator has a variable output attenuator, accurate in dB, the receiver can be operated with manual gain control (turn off agc) and the level controlled by the generator.

To measure Q of the parallel-resonant circuit, set the generator for maximum signal at the receiver and note the frequency. Then tune the generator to either side of this frequency, tracking the receiver tuning, until the signal is 3 dB less than peak and note the frequencies. Find the difference between the two 3 dB frequencies and divide into the peak frequency. The result is the total Q of the parallel circuit.

## what happens if the components aren't measured?

About the worst thing that can occur is increased insertion loss and a slight change in bandwidth. This assumes that the coils have been checked for inductance within 10 per cent of the calculated value at least with a dip-meter or inductance bridge.

The usual case is a Q reduction with toroids by turn spreading or bunching to trim it to right inductance. Another is close spacing of shielding on solenoid coils which lowers inductance as well as Q. Slug-tuned inductors can change Q by 50 per cent between slug in and slug out. Pot-core inductors are very good at frequencies below a 1 MHz but often the wrong core material is selected.

Beware of junkbox capacitors with partly legible markings. They may not be what you think they are; they may also be damaged. A multisection filter can still perform with one section grossly off resonance but the insertion loss will be dozens of dBs lower than you would expect.

A very common mistake goes all the way back to initial calculations. Double check your math. Check for section resonance at the center frequency. The best insurance against dropping decimal points is to use a pocket calculator.

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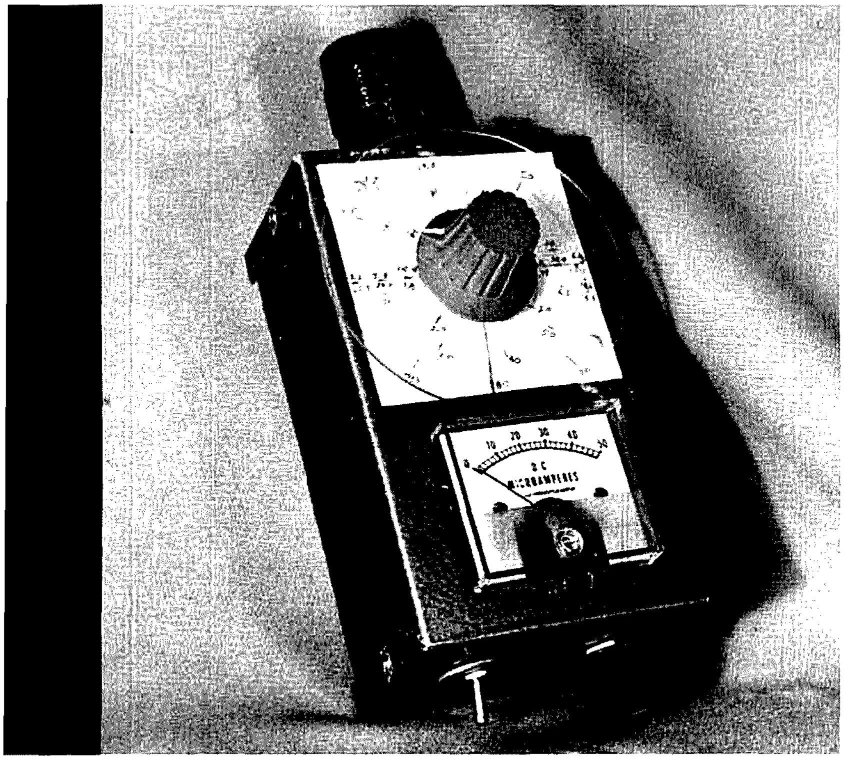
**ham radio**



# gate-dip meter

## that really dips

This gate dipper covers 1.8 - 150 MHz and can be built for less than $25

The dip oscillator described here is a solid-state version of the 6C4 vacuum-tube grid-dip meter in the 1957 *ARRL Handbook*. Unlike some of the single-fet Colpitts circuits described in the literature, this Hartley configuration really performs. In fact, a dip from about 50 - 20 microamps can be obtained on most bands when the dipper is held an inch (25mm) or so away from the resonant circuit under test. Furthermore, few false dips are encountered throughout the tuning range. The circuit is straightforward and the entire project can be completed in a weekend. If all new parts are used the cost should be less than $25, which isn't bad when compared with some of the commercial dip meters available.

The heart of the instrument is a Siliconix 2N5398 vhf uhf jfet (**fig. 1**); however, an MPF107 (2N5486) should work just about as well. Try several of one type and use the one that yields the highest off-resonance gate current. If you're not interested in the vhf capability, the MPF102 is by far the most economical device choice. Coil data appears in **table 1**.

### construction hints

A Minibox houses all parts with plenty of extra space inside as shown in the photo. Components are mounted on a one-inch-square (25mm) piece of copper-clad board, whose surface was sectionalized with a hacksaw. This board is located directly at the coil-socket terminals to minimize stray inductance; otherwise vhf operation will be limited.

At this point a few general comments are in order. My model contains a 50-pF tuning capacitor with one plate removed. If your tuning range is a little different, the reason can be attributed to a difference in tuning-capacitor value. The coil tap position becomes more critical at the higher frequencies. A point can be reached

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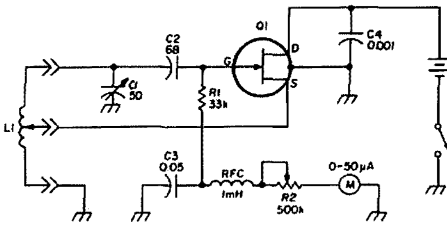


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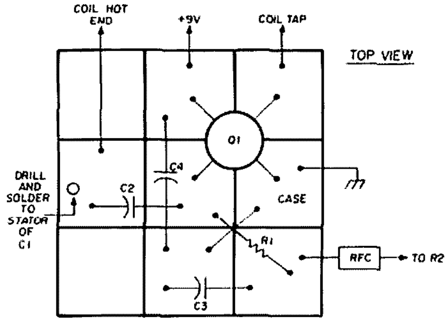



B

fig. 1. Schematic of the gate-dip oscillator, A, and suggested circuit-board layout, B (not to scale). C1, 50 pF (Hammarlund HF50 or equivalent). Q1 is a 2N5398, 2N5486, or MPF102 (see text). Coil data is shown in table 1. Power is supplied by a 9-volt transistor-radio battery.

where the dip is definitely most pronounced; therefore, some trial and error tinkering with the coil-tap point may be in order if you're really particular. The vhf coil is especially critical in this respect. Fortunately, the tap point is easily moved about on this hair-pin-shaped coil.

number of turns will change. Remember that inductance varies as the square of the number of turns.

## calibration

Calibration was accomplised by coupling the dipper to a coil (about 40 turns, 1 inch (25mm) diameter), which in turn was connected to a Hewlett-Packard 5300A counter for frequencies up to 50 MHz. A 5328Z counter was used for the higher frequencies. A receiver or another calibrated dip meter can also be used, of course.

Since you may prefer to design your own dials and chassis layout, I've left this part up to the reader. The photos show how I arranged the parts and made the

Open Minibox showing component layout. The 9-volt battery is mounted in the bottom half of the Minibox. One-hand operation, easy construction and low-cost parts make this a nice piece of test equipment.

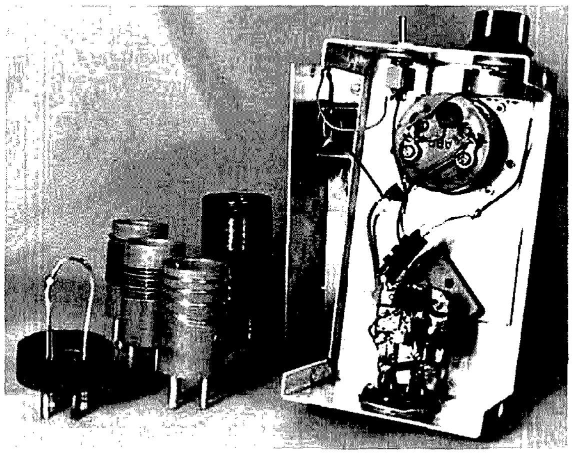
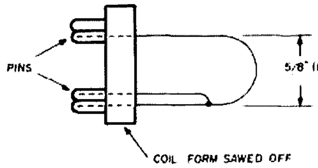


table 1. Coil data.

| frequency range (MHz) | no. turns | wire size AWG | wire size (mm) | winding length inches | winding length (mm) | tap* | coil diameter inches | coil diameter (mm) |
|---|---|---|---|---|---|---|---|---|
| 1.8 - 3.8 | 82 | 26 enamel | (0.4) | 1 9/16 | (40.0) | 12 | 1¼ | (32) |
| 3.6 - 7.3 | 29 | 26 enamel | (0.4) | 9/16 | (14.5) | 5 | 1¼ | (32) |
| 7.3 - 14.4 | 18 | 22 enamel | (0.6) | 3/4 | (19.0) | 3 | 1 | (25) |
| 14.4 - 32 | 7 | 22 enamel | (0.6) | 1/2 | (12.5) | 2 | 1 | (25) |
| 29 - 64 | 3½ | 18 tinned | (1.0) | 3/4 | (19.0) | 3/4 | 1 | (25) |

61 - 150 Hairpin of 16 no. AWG (1.3mm) wire, 5/8 inch (16mm) spacing, 2 3/8 inches (60mm) long including coil-form pins. Tapped at 2 inches (51mm) from ground end.




*Turns from ground-end. 1 inch (25mm) forms are Millen 45004 available from Burstein-Applebee

I made a coil-capacitor combination that resonated at about 100 MHz to adjust the vhf-coil tap for the best meter hull.

The total cost of this project can be reduced somewhat if you have some old four-prong vacuum tubes lying around. Their bases serve admirably as coil forms. The diameter of these forms is different from the one-inch (25mm) forms specified in the parts list, so the frequency dial. This dial arrangement, by the way, is a larger version of that described in the *ARRL Handbook* article and allows for convenient, one-hand operation.

## acknowledgement

I wish to extend my appreciation to Professor Jim Privitera for the photographs in this article.

**ham radio**

# toroid permeability meter

Complete design data and construction details for an instrument to determine the permeability of unknown toroid cores

**Small toroidal inductors wound** on cores of various permeabilities are being used today in great numbers. I obtained from a surplus source some toroid cores of different sizes and colors having an assortment of colored stripes and dots, plus house numbers of many digits, imprinted on them. To determine what kind of cores these are, in terms of permeability, you could wind a coil with a reasonable number of turns (about 10), and measure the resultant inductance using a bridge, Q meter, or grid-dip oscillator. Then the well-known formula for the inductance of a toroidal coil can be used to compute the permeability:

$$L = 0.004046\,\mu\,n^2\,h\,\log_{10}(b/a) \tag{1}$$

where $L$ = inductance ($\mu$H)
$\mu$ = permeability (non-dimensional)
$n$ = number of turns
$h$ = core height through hole* (cm)
$b$ = outside diameter (cm)
$a$ = inside diameter (cm)

However, it would be easier to place the unknown core into an instrument and directly determine permeability. Such a hands-off approach is what this article is all about: how to build a toroid-core permeability meter. The instrument described here uses as instrumentation only a grid-dip oscillator whose dip frequency signal is

*Height is the axial distance, not the distance measured along the coil radius, which is the thickness.

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monitored by a communications receiver with accurate frequency readout. The permeability meter has been designed with the average experimenter in mind and is built of surplus materials.

## background

In 1924 G. A. Kelsall[1,2] constructed a permeability meter for measuring the permeability of large iron toroidal windings used in power-line-frequency electrical machinery and low audio frequencies of that era. In 1953 his designs were modified by Haas, Edson, and others[3,4] of the National Bureau of Standards, for the construction of an rf toroid permeability meter. With simple modifications, both the Kelsall and Haas instruments are also capable of measuring the core dissipation factor (loss tangent) and permeability temperature coefficient as well as the permeability. Only the latter is of interest here. The references can be consulted for information on the other parameters.

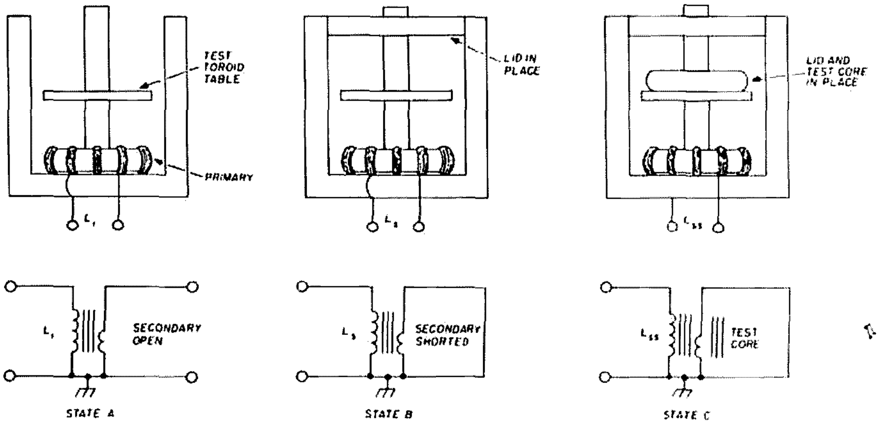
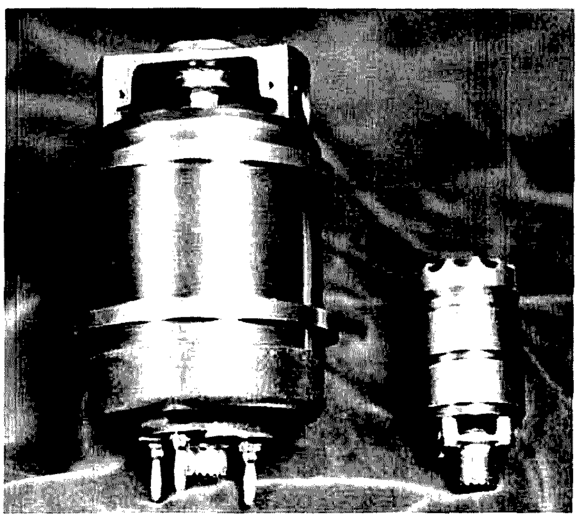



**fig. 1. Toroid permeability-meter states. The permeability meter is a step-down transformer consisting of a copper or brass cup and lid secondary, with a reference-toroid-core primary lying on the inside bottom of the cup.**



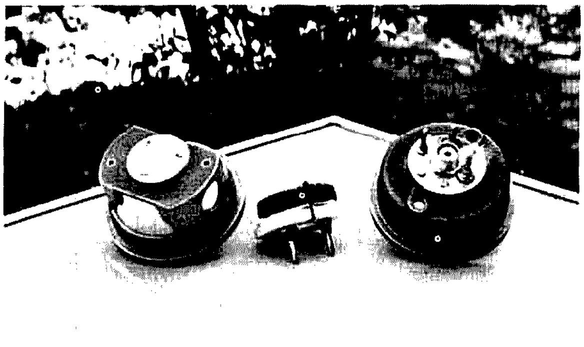
**Selsyn-case cup, left, and minicup permeability meters. On the selsyn-case cup note the large end of the upper half of the pop-up stem center conductor soldered to its shaft as well as to the top end bell. Banana plugs and coax-connector primary toroid-coil terminals protrude from bottom end bell.**

## design

The rf toroid permeability meter is essentially a step-down transformer of novel design. The primary winding is on a toroidal core, in this case, about 2 inches (50mm) in diameter with $n$ turns (the value of $n$ is discussed later). The primary winding is lying on, or plugged into, the inside bottom of an approximately 3-inch (77mm) diameter copper or brass cup. The cup contains a center conductor that passes through the toroid core primary and fastens to the cup lid, as in figs. 1 and 2. The cup and center conductor, with its lid in place, form a husky, three-dimensional, one-turn, shorted secondary winding. The lid shorts the secondary by connecting the center conductor to the cup wall. With the lid off, the secondary is open since that connection is broken.

The permeability meter provides an input inductance to be measured across the primary winding for each of three cup states, A, B, and C. In state A, fig. 1A, the cup lid is off (open secondary). In state B, fig. 1B, the cup lid is on (shorted secondary). In state C, fig. 1C, the cup lid is also on, but with the core whose permeability is sought in place on its table in the cup. The table is made of low-loss material, such as Teflon, and is positioned in the upper third of the cup. (This position is not critical.)

Fig. 1 shows the three states of the permeability meter when checking the permeability of an unknown core. $L_l$, the input inductance measured across the reference toroid coil primary with the secondary open, is the largest of the three inductances. $L_s$, the input inductance with the secondary shorted, is the smallest of the three inductances because it represents the primary with reduced inductance because of the shorted secondary negative reactance reflected back to the primary. $L_{ss}$, the input inductance with the secondary shorted, but with an increased secondary inductance because of the addition of the unknown core inductance in the second-



Exterior view of top and bottom cup end bells. At left is the top end bell showing circular brass plate covering the top central hole. In the center is the primary toroid coil on a Teflon base with Teflon sheet pad and banana-plug terminals. At right is the bottom end bell showing three banana plugs (one is a dummy) and coax-connector primary-coil outlet terminals. Notch in bottom brass plate is a quick fix for a too-large hole. The normal plate has an intact circumference.

ary magnetic circuit, is always intermediate in value between $L_l$ and $L_s$; *i.e.*, $L_l > L_{ss} > L_s$.

## testing a core

For those who have the instrumentation, the rest is easy. After the instrument is built, and with its lid on and a test core in place, (state C), $L_{ss}$ is measured on an inductance bridge. Then, with the lid off and no test core (state A), $L_l$ is similarly measured. With the lid on and no test core (state B), $L_s$ is also similarly measured. From the analysis in the appendix (eq. 1-11) the permeability of the unknown core is given by:

$$\mu - 1 = \left[0.004046\, h \log_{10} \frac{OD}{ID}\right]^{-1} \cdot \left[\frac{L_l^2}{n^2 (L_l - L_s)}\right] \cdot \left[\frac{L_{ss} - L_s}{L_l - L_{ss}}\right] \quad (2)$$

$OD$ = outside diameter of unknown core
$ID$ = inside diameter of unknown core
$L_l$ = input inductance in state A ($\mu$H)
$L_s$ = input inductance in state B ($\mu$H)
$L_{ss}$ = input inductance in state C ($\mu$H)

The other quantities are defined following eq. 1. Note that the middle factor of the expression above is a constant of the individual instrument. This will facilitate its calibration, as discussed later.

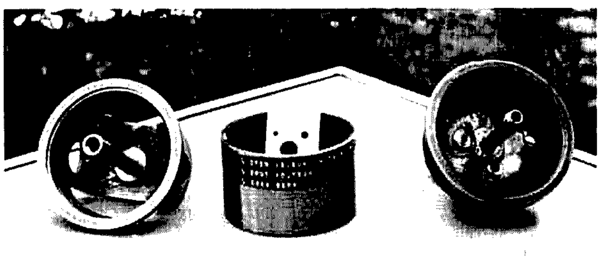
Using an electronic calculator with a log capability, permeability can be computed by plugging into eq. 2 the measured value of input inductance, number of turns of the primary inductance, and the unknown core dimensions.

Suppose you have no sophisticated instrumentation or calculator? The remainder of this article is devoted to the problem of determining the permeability of unknown cores using a permeability meter built from surplus and other readily available low-priced materials. The permeability meter operates in conjunction with a grid-dip oscillator and a communications receiver with an accurate dial readout (3-figure accuracy below 10 MHz). The unknown permeability is obtained from a nomogram, together with the grid-dip oscillator frequency-dip data and the unknown core dimensions.

## construction

As the accompanying figures show, the essential requirement is a copper or brass cup with a removable threaded or beveled lid that also holds the cup center conductor. For example, a center conductor made from 1/4-inch (6.5mm) threaded brass rod could be used with the cup. This, of course, would limit core sizes to IDs greater than 1/4 inch (6.5mm) so that the cores can slip over the center conductor when placed into the cup. Cup diameter should be large enough to accommodate the largest cores anticipated by the user. A cup diameter of 3½ inches (89.3mm) is used here. Cup height-to-diameter ratio isn't critical but should be close to unity (height $\approx$ diameter).

Two banana plugs and jacks provide plug-in capability inside the cup for the primary toroid coil and connections to the outside. The plugs and jacks are soldered back-to-back after removing the threaded portion of the banana plug. One plug is insulated from the cup bottom with Teflon shoulder washers, and one is in contact with the cup (see fig. 2). A third dummy banana plug is also provided to extend plug-in versatility, as explained later. This completes the instrument except for the primary coil (described below). I added an SO-239 coax connector to connect other test equipment, but this connector isn't necessary. The primary coil is mounted on a base of low-loss material with banana plugs, as shown in fig. 2 and the photos.



Interior view of top and bottom end bells. At left is the interior of the top end bell integral (soldered) with the upper half of the pop-up stem center conductor. The center is the Teflon block table leaning on the wall of the central brass shell of the instrument. Right shows the interior of the bottom end bell with the jack ends of the banana-plug and jack primary-coil connectors. Note window of large end of pop-up stem to accommodate jumper wire (with sleeving) from coax center conductor to insulated banana-plug/jack. The lower plug/jack is grounded to the cup bottom.

Making the case. Many inexpensive surplus selsyn motors are available that are about 3 inches (77mm) in diameter by about 5 inches (128mm) high. These are typified by a 110 V, 60 Hz, 5 ampere, type M unit. They have a brass

case and heavy flanges bearing manufacturing dates between 1943-1956.* All we need is the brass case. Remove the electrical terminals with a screwdriver, then remove the field coils using a claw hammer. Use pliers to remove the thin field-coil laminations, two or three at a time. Easy does it. The case will now come apart in three pieces: top and bottom end bells and center shell.

At each end of the bare case is a circular brass plate covering a 7/8-inch (22mm) hole. It turns out that a brass lawn sprinkler pop-up stem† found in hardware stores has a large end that just fits into the selsyn case end holes. Two of these pop-up stems screwed back-to-back also just fit into the selsyn case to make the permeability meter center conductor. The pop-up stems have a partially threaded bushing, one of which should be threaded all the way through so that the center conductor can be screwed and unscrewed as the instrument lid is installed and removed during use.

Soldering the large end of each pop-up stem onto its shaft, as well as into the end bells of the selsyn case, can be done easily with a small Butane torch. First, however, the bottom end bell of the selsyn case should be drilled with three holes through the bottom brass plate for the banana plugs and jacks for the primary toroid coil outlet, as mentioned previously. The three holes form a right triangle on the circular brass plate, with the insulated hole at the right-angle corner and the two non-insulated holes at the other two corners. The distances from the insulated hole to the noninsulated holes, along the short sides of the triangle, are 1 inch and 3/4 inches (25.5 and 19mm) respectively. These dimensions correspond to most inductance-bridge and Q-meter-terminal spacings.** These, and other dimensions, are shown in **fig. 2**.

The bottom end bell central hole will take a nut-type coax connector. This type of connector mounts with two large mounting nuts in a 5/8 inch (16mm) hole, instead of the usual square SO-239 mounting. The connector is then mounted in the circular brass end plate, with a 5/8 inch (16mm) hole punched out. This assembly is screwed back into the bottom end of the cup, as shown in **fig. 2**. Also, before soldering the pop-up stem into the bottom (done with the brass plate in, but with its accessories removed), a small window must be notched out of the large end of the stem to provide access to the SO-239 connector center conductor. The latter is jumpered with a short piece of sleeve-covered wire to the insulated banana jack inside the cup. This is the only wiring in the instrument. This connection is shown in

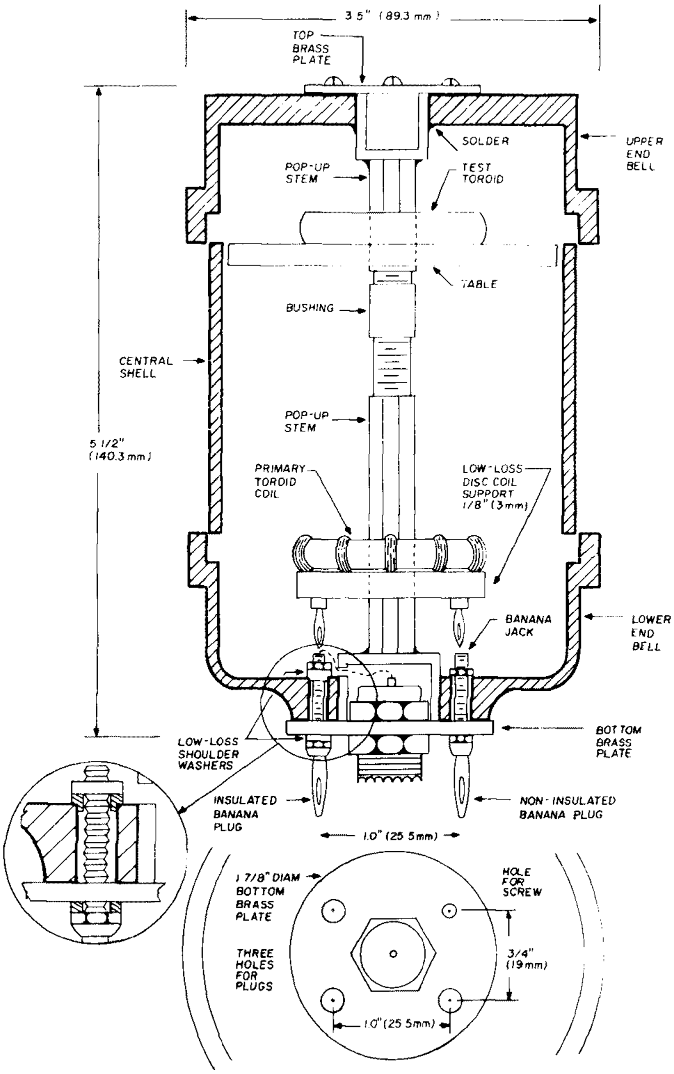



**fig. 2. Toroid permeability meter built from a surplus selsyn-motor case and a lawn-sprinkler pop-up stem.**

**fig. 2**. It is made after soldering the pop-up stem in place.

**Primary coil.** The primary coil consists of 31.6 turns (so that $n^2$ will equal 1000), no. 24 AWG (0.5mm) wire, which is equally spaced around an Arnold Engineering Company toroid core.* Any other core of equivalent size, quality, and permeability may be used providing it meets the constraints discussed later on using the grid-dip oscillator/amateur receiver combination to determine unknown core permeability.

From **fig. 2** and the photographs it is seen that the cup lid; *i.e.*, the upper end bell and the upper half of the

*Manufactured by Henschel Corporation, Amesbury, Massachusetts.

†Manufactured by Champion Brass Company, 1460 Naud Street, Los Angeles, California 90012. Available from Sears.

** Either of the two noninsulated plugs is a dummy, since the primary coil needs but two jacks to be seated inside the cup bottom. The dummy plug is included so that all three plugs, two at a time, accommodate most bridges. Of course, the insulated plug is one of the two used.

*Arnold Engineering Company, P.O. Box G, Marengo, Illinois 60152. Dimensions are 1.875 inch (48mm) OD; 1.375 inch (35mm) ID; and 0.375 inch (9.6mm) in height. Permeability, $\mu$, is 125. Part no. is D18002/AM-12.

center conductor, are now integral and comprise the upper portion of the instrument. The middle portion consists of the thin brass shell, while the bottom end bell contains the lower half of the center conductor and the banana plugs and jacks for the primary coil. The central shell is a press-fit into the bottom end bell of the cup. It is also a press-fit into the top end bell as the center conductor is screwed together during test of a core. With a low-loss (Teflon) table or support strip drilled and tapped for the center conductor, where the table is positioned to occupy the upper third of the cup, the assembly is complete.

**Minicup permeability meter.** This instrument, which is also shown in the photos, is about salt shaker size and is used for measuring the permeability of tiny cores. It is made from scrap brass and is 1.5 inch (38mm) OD, 1.25 inch (32mm) ID, and 1.75 inch (45mm) long. It is mounted on an SO-239 chassis-mount connector. The primary coil is 31.5 turns of no. 28 AWG (0.3mm) wire spaced on an Amidon T-94-2 (red) core. The primary leads feed through a cup bottom hole; one is soldered to the SO-239 center conductor and the other is grounded to the cup. The center conductor is a 6-32 (M3-5) machine screw. The lid is made from a piece of cylindrical brass scrap.

In the appendix, **eqs. 1-13** and **1-14** give another version of the formula for permeability shown in **eq. 2**. This formula is:

$$\mu - 1 = \frac{k}{h \log_{10}(OD/ID)} \frac{(f^2_{HI} - f^2_{cor})}{(f^2_{cor} - f^2_{LO})}$$

$$f_{HI} > f_{cor} > f_{LO} \tag{3}$$

$k$ = a constant of the instrument obtained by calibration as discussed below

$f_{HI}$ = the upper grid-dip oscillator frequency measured with no core inside the cup, and with the lid on. This corresponds to the $L_s$ state **B** of the instrument

$f_{LO}$ = the lower grid-dip oscillator frequency measured with no core inside the cup, and with the lid off. This corresponds to the $L_l$ state **A** of the instrument

$f_{cor}$ = the intermediate grid-dip oscillator frequency measured with the unknown core inside the cup, positioned on its table, and with the lid on. This corresponds to the $L_{ss}$ state **C**

The other quantities in **eq. 3** are defined following **eqs. 1** and **2**. Now, the analysis in the appendix (**eq. 1-15**) modified as above for the constant, $k$, gives for this constant and the secondary inductance, $L_{coax}$,

$$k = \frac{L_{coax} - L_l/n^2}{0.004046} \tag{4}$$

and

$$L_{coax} = 0.004046\, H \log_{10} (D_{cup}/D_{cen})$$

$L_{coax}$ = secondary inductance ($\mu$H)
$D_{cup}$ = cup inner diameter (cm)
$D_{cen}$ = center conductor outer diameter (cm)
$H$ = cup height (cm)

A slight digression is appropriate at this point concerning the structure of the toroid inductance formulas. First, **eq. 1** for the inductance of a toroid contains three main factors:

1. The permeability, $\mu$, which is a measure of the inductance enhancement of the coil over its free (air-wound) space version.

2. The square of the number of turns, $n^2$.

3. The factor $0.004046\, h \log_{10} (OD/ID)$, which is the inductance in $\mu$H of the free space that the toroidal core volume occupies.

Now, the shorted cup secondary surrounds a toroidal volume of free space and thus has an equivalent inductance, $L_{coax}$, as in **eq. 4**. Note that this formula has the same structure as the free-space inductance of the unknown toroid, **eq. 1**, as it should. $L_{coax}$ also means that it is, at the same time, the inductance of a short, fat section of air-dielectric coax of length $H$ and diameters $D_{cen}$ and $D_{cup}$, which also happens to surround a toroid volume of free space.

Second, the formulas above for permeability hold ideally *only* when the primary leakage inductance is zero or negligibly small, as discussed in reference 3. However, with the number of bolt holes and cast cutouts in the surplus selsyn case, the primary leakage inductance can be appreciable since some of the primary flux that threads these holes is sure not to link the secondary, which contributes to the leakage flux and thus to the leakage inductance. Hence, the above constant, $k$, is not given accurately by **eq. 4**. Instead, $k$ will be found by calibration with a number of cores of known permeabil-

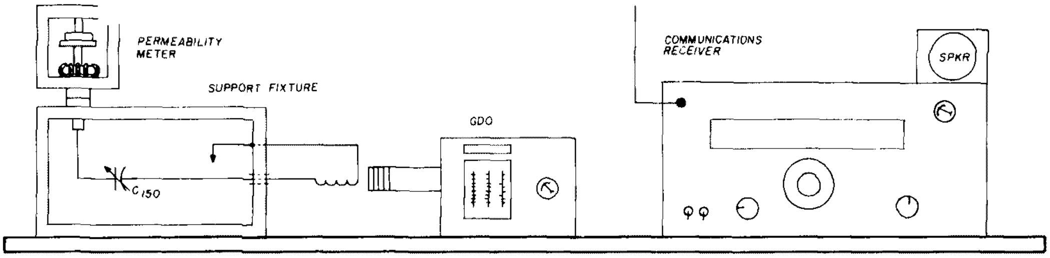


**fig. 3. Instrumentation for obtaining input to the nomograms.**

table 1. Toroid permeability meter calibration data obtained with primary toroid coil in the 7-8 MHz range.

| serial number | core type (1) | known perm, $\mu$ | freq (2) (MHz) | OD inches | (mm) | ID inches | (mm) | height inches | (mm) | $f_{LO}$(3) (MHz) | $f_{HI}$(3) (MHz) | $f_{corr}$(3) (MHz) | $k$(4) | calculated permeability |
|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|
| 1 | T94-10 (blk) | 6 | 60-150 | 0.942 | (24) | 0.560 | (14) | 0.312 | ( 8) | 7.090 | 7.984 | 7.903 | 8.47 | 6.94 |
| 2 | T94-6 (yel) | 8 | 10-90 | 0.942 | (24) | 0.560 | (14) | 0.312 | ( 8) | 7.087 | 7.993 | 7.884 | 8.64 | 9.10 |
| 3 | T94-2 (red) | 10 | 1-30 | 0.942 | (24) | 0.560 | (14) | 0.312 | ( 8) | 7.086 | 7.982 | 7.862 | 9.78 | 10.2 |
| 4 | T94-3 (grey) | 35 | 0.05-0.5 | 0.942 | (24) | 0.560 | (14) | 0.312 | ( 8) | 7.089 | 7.989 | 7.696 | 11.93 | 29.5 |
| 5 | FT-114-63 | 40 | 1.5-25 | 1.142 | (29) | 0.748 | (19) | 0.295 | (7.5) | 7.087 | 7.980 | 7.613 | 7.26 | 54.7 |
| 6 | T94-41 (grn) | 75 | 0.001-0.10 | 0.942 | (24) | 0.560 | (14) | 0.312 | ( 8) | 7.077 | 8.004 | 7.894 | 0.81 | 9.07 |
| 7 | FT-114-61 | 125 | 0.2-10 | 1.142 | (29) | 0.748 | (19) | 0.295 | (7.5) | 7.087 | 7.984 | 7.545 | 16.75 | 75 |
| 8 | FT-114-43 | 950 | 60-200 | 1.142 | (29) | 0.748 | (19) | 0.295 | (7.5) | 7.085 | 7.986 | 7.153 | 9.95 | 955 |
| 9 | FT-114-72 | 2000 | 0.001-1 | 1.142 | (29) | 0.748 | (19) | 0.295 | (7.5) | 7.065 | 7.995 | 7.091 | 13.35 | 1498 |
| 10 | Arn. Eng AM-12 | 125 | 0.2-20 | 1.875 | (48) | 1.375 | (35) | 0.375 | (9.6) | 7.089 | 7.992 | 7.646 | 24.08 | 111 |

notes:

(1) Amidon Associates nomenclature except core 10. T prefix indicates powdered iron; FT prefix indicates ferrite.

(2) Manufacturer's recommended frequencies.

(3) Meaured using a 1947 Millen grid-dip oscillator and surplus ARR-41 receiver.

(4) Calculated using **eq. 3** with frequency measurements described in text and known permeabilities. $k_{AV} \approx 10$.

ity. Again, $k$ could ideally be calculated from **eq. 4**, and no calibration would be necessary if, a) the leakage inductance were negligible, and b) the correct cylindrically equivalent physical dimensions; *i.e.*, $H$, $D_{cup}$, $D_{cen}$; of the selsyn cup, were known.

For calibration as well as for normal use, the permeability meter is connected in series with a variable capacitor and a 2- or 3-turn link and coupled to a grid-dip oscillator, as in **fig. 3**. A communications receiver with a reasonably accurate dial readout is used to pick up the grid-dip oscillator signal. A small antenna about a foot (30cm) long on the receiver is sufficient. This combination of a common grid-dip oscillator and a medium-quality communications receiver should provide a frequency readout to three figures. Amateur-band-only receivers can usually provide better accuracy. However, the permeability meter must be matched in physical construction so that the frequency spread, $\Delta f = f_{HI} - f_{LO}$, doesn't span more than 500 kHz, which is the usual vfo range on most ham-band-only receivers. This constraint is discussed in detail in a later part of the article.

## calibration data

For ten cores of known permeability, claimed to be correct within 5-10 per cent, **table 1** gives the grid-dip oscillator/receiver data I used to calibrate the permeability meter shown in the photos. Using an average $k$ of about 10 from the computations described above, unknown toroid permeabilities should be about ±10-20 per cent of their actual values. This can be seen by examining **table 1** data and by computing the deviation of $k$ from the average $k$ for each core (except core no. 6). Other statistics, such as the mean-square deviation of the permeabilities for a series of readings on each core, are left as an exercise for the interested reader.

Core no. 6 gave a computed permeability way out of the ballpark compared with its claimed value. This is probably because it was measured in a frequency range of 7-8 MHz, which is remote from its intended use range (0.001-0.1 MHz). Data from this core was therefore not used in computing the average $k$.

## using the nomogram

As mentioned earlier, for those who don't have calculators with a logarithm capability, or inductance bridges or Q Meters to obtain the unknown toroid permeability, **figs. 4** and **5** comprise a nomogram to use for this purpose. The unknown toroid dimensions and grid-dip oscillator frequency-dip readings are required as input to the nomogram. **Eq. 3** for the permeability, $\mu$, slightly modified as in **eq. 5**, is the basic equation from which the nomogram is constructed. If your permeability meter is about the same size as mine, the constant, $k$, will be about 10, as seen from **table 1**. For other sizes of instruments, and thus other values of $k$, corresponding adjustments can be made in the nomogram as explained below.

A second parameter of the instrument is the ratio $f_{LO}/f_{HI}$, called $a$, of the instrument. It appears when **eq. 3** for the permeability, $\mu$, is rewritten as

$$\mu - 1 = \left[\frac{k}{h \log_{10}(OD/ID)}\right] \bullet \left[(1 - f^2)/(f^2 - a^2)\right] \quad \textbf{(5)}$$

where $f = f_{cor}/f_{HI}$, and the other quantities are as defined earlier. The nomogram in **figs. 4** and **5** will yield permeabilities of unknown toroid cores for corresponding sets of $h$, $OD$, $ID$, $a$, and $f$ values.

**Directions for use.** The following directions for using the nomogram are augmented later by a step-by-step example. For now, noting the sketches at the bottom of **fig. 5**, the directions are as follows.

1. Beginning on **fig. 4** nomogram, with a transparent straightedge, find the value of the unknown toroid core $OD$ on line 1 and its $ID$ on line 2. Span these values and lightly mark where the straightedge crosses line 3.

2. Span the straightedge from the line 3 mark to the value of height on the toroid height line 4, then lightly mark where it crosses diagonal line 5. Note the value of the latter mark as it will be needed later.

3. Referring to the fig. 5 nomogram, from right to left, and knowing the values of $a$ and $f$ from the grid-dip oscillator frequency measurements, span from the $a$ value on line 6 to the $f$ value on line 7. Lightly mark where the straightedge crosses line 8.*

4. Spanning from the mark on line 8 to $f$ on the $f$ line (line 9), lightly mark where the diagonal line 10 is crossed.

5. Noting the value on line 10, find the same value on the diagonal line 10 on the fig. 4 nomogram.

6. Span from the fig. 4 line 10 value to the diagonal line 5 value on the same figure. Mark where the straightedge crosses the permeability line 11.

What you have really found on line 11 of fig. 4 is $\mu - 1$ but for all except the smallest values of permeability, the above line 11 mark is essentially the permeability (within the accuracy of the instrument and the nomogram procedure). For small permeability values near one, merely add one to the value found on line 11.

For those whose permeability meter dimensions are radically different from this selsyn-case instrument, the line 11 scale should be interpreted as $10(\mu - 1)/k$, where $k$ is the constant for your instrument. With a number of cores of known permeability, you will be able to determine a $k$ for your instrument such that the permeability line 11 will fit the known toroid permeabilities. Then you can relabel the scale on line 11 accordingly. This is the reason for the numbers less than one that are on the line 11 scale. For the minicup permeability meter, $k$ is 3.5, while $k$ for the selsyn-case cup is 10.

Examples. Here are examples for obtaining the permeability of three unknown toroids. The first example involves a toroid from my batch of surplus cores. The core is medium gray in color with the black numerals 262-55582-A2 around the edge. Its dimensions are 1.375 inch (35mm) OD, 0.94 inch (24mm) ID, and height is 0.35 inch (9mm). Grid-dip oscillator/receiver frequency-dip measurements were: $f_{LO}$ = 7.07 MHz, $f_{HI}$ = 7.98 MHz, and $f_{cor}$ = 7.46 MHz. Immediately, $a$ is 0.885 and $f$ is 0.934. Using a triangle for a straightedge, the procedure is:

1. On fig. 4 span from an OD of 1.375 inch (35mm) on line 1 to an ID of 0.94 inch (24mm) on line 2. This yields 0.15 on line 3.

2. Span from 0.15 on line 3 to the core height of 0.35 inch (9mm) on line 4. Note that the straightedge crosses line 5 at 0.14. Note this number as it will be used later.

3. Go to the nomogram in fig. 5. Span the straightedge

*On lines 6 through 9 of fig. 5, two quantities represent each line. In each case, either one can be used, whichever is convenient, as the lines are scaled accordingly.

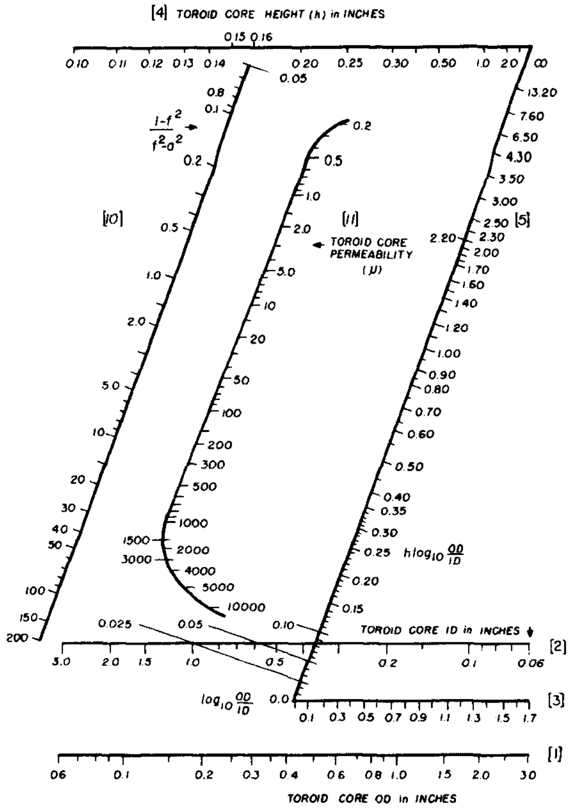



fig. 4. Nomogram for toroid dimensions and core permeability.

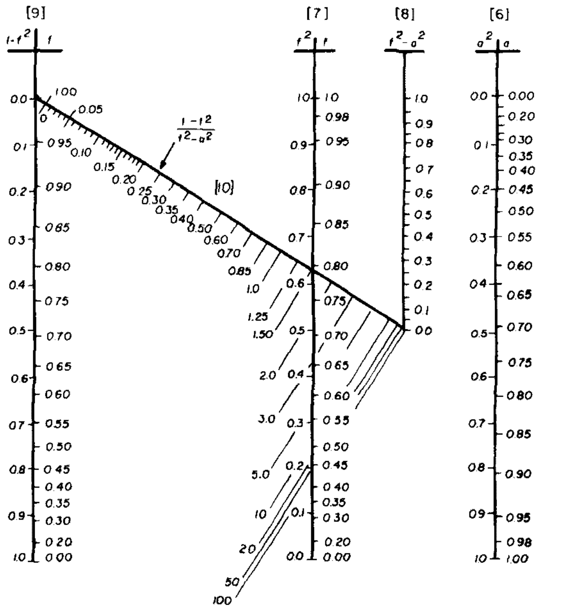
from $a$ = 0.885 on line 6 to $f$ = 0.934 on line 7. The straightedge crosses line 8 at about 0.10.

4. Span from 0.10 on line 8 to $f$ = 0.934 on line 9. The straightedge crosses diagonal line 10 at 1.5.

5. Go back to the fig. 4 nomogram. Find 1.5 on diagonal line 10. With 1.5 on diagonal line 10 and 0.14 on diagonal line 5, which you obtained from step 2, note that the straightedge span between these two numbers crosses the diagonal line 11 at about 90, which is the permeability of this core.

So the permeability of this toroid core is 90 ±10-20 per cent. When I obtained the grid-dip oscillator frequency dip data with the core inside the cup and lid on to get $f$, the dip was rather broad compared with dips without the core, and the corresponding grid-dip oscillator dial feel was sluggish. This implies that the toroid $Q$ is not very high, at least in the range between 7-8 MHz, as compared with the usual toroid rf core. I would estimate the $Q$ in this case, from the reciprocal of the fractional frequency spread, at about 10.

The conclusions are that this toroid core has a moderately high permeability of about 100 and probably was used in an af or i-f application. The surface, after some paint was scraped away, had the dull gray color and texture associated with powdered iron. This is meant in the sense that powdered iron is supposed to be dull gray in color and somewhat frangible, while ferrite is supposed to be darker, shinier,




THE SKETCHES BELOW ARE A PICTORIAL DESCRIPTION OF STEP-BY-STEP USE OF THE NOMOGRAMS IN FIGS. 4 AND 5.

FIG 4 NOMOGRAM

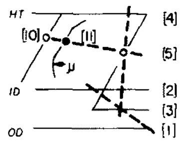



FIG 5 NOMOGRAM

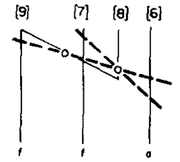



**fig. 5. Nomogram for grid-dip oscillator/communications receiver frequency data.**

and harder. Lending some credence to this rationale is that the core was roughly textured, which smacks of hurried manufacture.

For a second example, two toroid cores from my surplus batch are tan colored and both are imprinted with the red-colored edge number 2090257, surrounded by two red squares and a tiny five-pointed star with what looks like "AL" in its center. On the opposite edges are dark brown and dark red dabs of paint. Both have ODs of 0.8 inch (20.5mm), IDs of 0.5 inch (12.5mm), and heights of 0.25 inch (6.5mm).

The grid-dip oscillator/receiver measurements for the dark brown and dark red cores gave $a = 0.863$; $0.863$ and $f = 0.943$; $0.945$ respectively. Running them through the nomograms in the manner described above yielded permeabilities of 60 for the dark brown and 55 for the dark red core. (Their permeabilities are probably identical.) Scraping the paint on each revealed a shiny solder-like surface coating, and beneath that, the dull gray powdered-iron appearance. They seem to have been carefully made, as all edges had the appearance of having been tumbled before painting. The grid-dip oscillator dips seemed sharp, so their *Qs* are reasonably high at rf – probably a high i-f, low rf application.

A third example is an all-black, unpainted core with no markings of any kind. Its OD is 0.875 inch (22mm), ID is 0.56 inch (14.5mm), and height is 0.25 inch (6.5mm). Grid-dip oscillator/receiver data was $a = 0.863$ and $f = 0.876$. The nomograms yielded a permeability of about 850. The grid-dip oscillator dip was sharp, so the Q is high at rf. The core is well made of what seems to be hard ferrite material.

## grid-dip oscillator/ham-band-only receiver combination

As mentioned earlier, the grid-dip oscillator frequency spread, $\Delta f = f_{HI} - f_{LO}$, for this selsyn-case instrument is about 1 MHz. This spread is about right for my surplus ARR-41 receiver, which tunes 1 MHz with each band change. For amateur-band-only receivers, the usual vfo range is 500 kHz per band, with a frequency readout-accuracy of three or four places below 10 MHz (to 1 kHz) being commonplace.

To exploit this readout accuracy, which means to make the $\Delta f$ of the permeability meter 500 kHz or less to fit the vfo range, implies the existence of a design criterion of some sort. From the appendix (**eq. 1-18**) such a relationship is obtained between the primary toroid inductance, $L_l$, the secondary cup inductance, $L_{coax}$, and the fractional frequency spread, $\Delta f/f_{LO}$. It is

$$L_l/n^2 L_{coax} = 1 - \frac{1}{(1 + \Delta f/f_{LO})^2} \qquad (6)$$

To illustrate this criterion, consider two examples.

**1.** Suppose you'd like to use, for grid-dip-oscillator measuring purposes, the 500 kHz of the 20-meter position of a ham-band-only receiver. Then the corresponding fractional frequency spread is

$$\Delta f/f_{LO} = 0.5\ MHz/14\ MHz = 1/28 \qquad (7)$$

Note that, from **eq. 1**, $L_l/n^2 = \mu_l L_{la}$ as well, where $\mu_l$ is the permeability of the primary core, and the inductance of the volume occupied by this is $L_{la} = 0.004046\ h_l \log_{10}(OD)_l/(ID)_l$ in $\mu$H. Subscripts $l$ refer to the parameters of coil $L_l$.

Now from **eq. 6**, with $\Delta f/f_{LO} = 1/28$, yields

$$L_l/n^2\ L_{coax} = \mu_l\ L_{la}/L_{coax} = 1/14 \qquad (8)$$

This means that the ratio of $L_l/n^2$ or $\mu_l\ L_{la}$ to $L_{coax}$ must be 1/14 so that the instrument will accommodate the grid-dip oscillator frequency spread of 500 kHz.

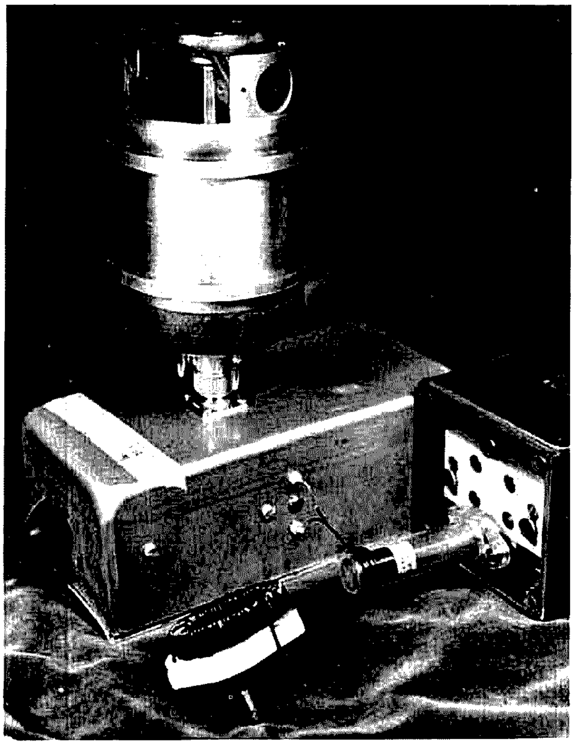
If your instrument is about the same size as mine, then $L_{coax}$ will be about 56 nH.* Then **eq. 8** says that $L_l/n^2 = 4\ nH$ is required for the 500 kHz of the 20-meter band on the ham-band-only receiver. Or, saying the same thing, $\mu_l\ L_{la} = 4\ nH$ must also hold by virtue of the preceding discussion. Then the core dimensions, number of turns, and permeability are chosen to satisfy both of these two expressions – plus the fact that

*The values of 13 $\mu$H for $L_l$ and 10 $\mu$H for $L_s$ for this instrument were measured on a Boonton 260 Q meter. From the appendix (**eq. 1-12**), $L_{coax}$ is 56 nH using these values and $n^2$ is 1000.

inductance must be compatible with the capacitor, C, in the grid-dip oscillator loop to resonate in the 20-meter band. The latter point is discussed below.

These relationships are not required to hold exactly. This is meant in the sense that the capacitor, C, can be used to partially compress or expand the frequency spread, $\Delta f$, to compensate for discrepancies between theory and practice. However, this is limited by the maximum and minimum values of capacitor C.

2. Another criterion must be satisfied; namely, the capacitor, C, must tune the primary inductance, $L_l$, in the frequency range of interest. The following example illustrates this point as well.



Selsyn-case cup permeability meter in place on a support fixture, which also supports the grid-dip oscillator link. Knob on the fixture varies the loop capacitor, C. The grid-dip oscillator is next to and slightly to the rear of the support fixture. The toroid primary coil is in the foreground. It is laced to Teflon-sheet padding on a Teflon base. Banana plugs allow coil to be plugged into bottom of cup. The coil consists of 100 turns of no. 24 AWG (0.5mm) wire on an Arnold Engineering type AM-12 core (core no. 10 in table 1).

Assume I want to convert my selsyn-case instrument to work with the grid-dip oscillator in the first 200 kHz of 20 meters with a Collins 75S-series receiver. The corresponding fractional frequency spread is $\Delta f/f_{LO} = 0.2\ MHz/14\ MHz = 1/70$. From eq. 6, my new $L_l/n^2L_{coax} = 1/35$. My present value is 3/13, as calculated from eq. 6 with $\Delta f/f_{LO} = 1\ MHz/7\ MHz = 1/7$. This means I must reduce 3/13 by a factor of about 8 to obtain 1/35. I certainly am not going to change the cup dimensions; *i.e.*, $L_{coax}$, so I must alter the primary toroid coil to achieve this change. This means I must simultaneously satisfy both the fractional-frequency-spread relationship, $L_l/n^2L_{coax} = \mu_l L_{la}/L_{coax} = 1/35$, which characterizes the primary coil physical dimensions and the resonant-frequency relationship, $L_lC = 1/4\pi^2 \cdot (14\ MHz)^2$, which fixes the number of turns, $n$. When all of this is amalgamated, using the $L_{coax}$ value of 56 nH, the implication is that the two relationships, $\mu_l\, h_l\, log_{10}\, (OD)_l/(ID)_l = 0.4$, and $n^2 = 80{,}000/C$ (capacitance in pF) hold at the same time.

Choosing $C = 50\ pF$ yields $n = 40$ for the number of turns required on the primary inductance. For the former relationship, if I were to use the same physical dimensions (for $h$, $OD$, and $ID$) for this new primary core as for the old core (core 10 in table 1), then its permeability works out to about 3. I then start looking through the catalogs for such a core.

In neither of the examples above would the nomogram require modification within the accuracies discussed earlier. This is true because (from eq. 1-16 in the appendix) the instrument constant, $k$, as expressed as $k = (L_{coax} - L_l/n^2)/4$ (inductances now in nH) will still remain the same.

To illustrate, note that the present $L_{coax}$ is 56 nH. With $L_l/n^2 = 13\ nH$, the corresponding $k$ is about 11 from the above expression. Now compare example 1, where $L_l/n^2 = 4\ nH$, which then gives about 13 for the above $k$. The same holds for example 2, where $L_l/n^2 = 1.6\ nH$, yielding only a slightly different $k$ of 13.5.

If the fractional frequency spread, $\Delta f/f_{LO}$, is always somewhat less than unity, then $k$ won't change very much. This is also shown in the appendix. However, the other instrument constant, $a$, will change; but once computed, $a$ is easily dealt with when going through the nomogram procedure.

## future work

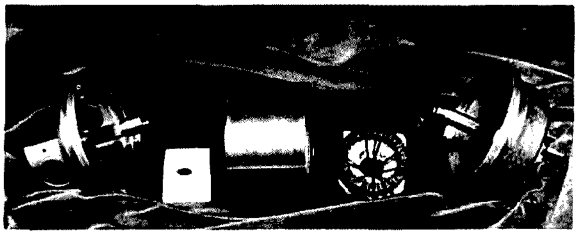
A number of areas of improvement in construction of the permeability meter suggest themselves. The selsyn case for the cup and the pop-up stems, whose dimensions were just right, were admittedly a coincidence. Many other copper or brass cup-like objects of this size are available. Hence, with a little extra scrutiny in this direction, you should be readily rewarded by a copper or brass cylinder of the right dimensions, even with top and bottom covers. An obvious approach, if facilities are available, is to build a cup from brass, with a bolted bottom and screw-on top and silver flashed circular fingers to receive the center conductor when the lid is on. This was done at the Bureau of Standards, as described in references 3 and 4.

A second area is that of achieving a more accurate frequency readout than can be obtained in the average communications receiver, by using an amateur-band receiver. Most receivers of this type have 3- or 4-place accuracy or better, but they span 500 kHz per band so that you cannot switch to the adjacent band segment to get the remainder of the frequency spread if the latter is

greater than 500 kHz. To make the grid-dip oscillator measured frequency span 500 kHz or less, attention must be given to the physical dimensions of the instrument design.

Discussion is continued here to point out the matching of the cup dimensions to the grid-dip oscillator/receiver combination for optimum operation. For example, optimum operation criteria could include cup design specifications for increasing the frequency spread, $\Delta f$, of a continuous-coverage quality receiver to obtain maximum accuracy of grid-dip oscillator frequency data.

As a case in point, it is shown in **appendix 1**, last paragraph, using a criterion of uniform measurement accuracy from low to high permeabilities, that the optimum $a^2$ in the formula for the unknown inductance (**appendix 1 eq. 1-19**), namely



**Exploded view of the selsyn-case cup premeability meter. Top end bell is at left with bushing screwed onto upper half of pop-up stem center conductor. Next is the Teflon-block table to support unknown toroid core. In the center is a thin brass shell, which press fits into the top and bottom end bells. Next is a primary coil of 31.5 turns of no. 24 AWG (0.5mm) wire on an AM-12 core (core no. 10 in table 1). At right is the bottom end bell integral with the lower half of the center conductor. Banana plugs and coax connector protrude to the right of cup bottom end bell.**

$$L_\mu = \left[L_{coax}\right]\left[\frac{a^2(1-f^2)}{(f^2-a^2)}\right] \qquad (8)$$

should equal 1/2. That is, $a^2 = (f_{LO}/f_{HI})^2 = ½$. This, then, implies that, for optimality in this sense, the following should hold:

1. The fractional-frequency spread, $\Delta f/f_{LO} = \sqrt{2} - 1 = 0.414$, or $f_{HI} = \sqrt{2}\, f_{LO}$, so that the optimal upper grid-dip oscillator frequency should be 41.4 per cent greater than the lower grid-dip oscillator frequency. In my case, if I insist on $f_{LO} = 7\ MHz$, then $f_{HI} = 10\ MHz$.

For an amateur-band receiver (or transceiver), with a vfo frequency spread of $\Delta f = 500\ kHz$, the corresponding $f_{LO} = \Delta f/0.414 = 1.2\ MHz$, which is smack in the middle of the broadcast band. So the next best thing is to use 80 meters (3.5-4.0 MHz), or 160 meters, in the off chance that 500 kHz is available on your receiver.

2. From the above, the instrument design relationship is now such that $L_l/n^2\, L_{coax} = 1 - a^2 = ½$, as well. This also means that $\mu_l L_{l_a} = ½ L_{coax}$; *i.e.*, the primary toroid geometrical or free-space inductance should equal half that of the secondary or cup geometrical inductance.

A third area of improvement is the construction of a more efficient nomograph, perhaps one with fewer scales. Or you could resort to graphical plotting procedures, which would amount to the same complexity.

A fourth area of improvement might be a direct-reading permeability meter. A tertiary winding could be introduced within the cup beneath the table. This winding would surround the center conductor to sample the cup magnetic-field strength. The tertiary winding terminals could easily be brought out through the cup wall. The internal cup field would vary in strength with cores of varying permeability. Diode rectification of the corresponding probe current could be used to deflect a permeability-calibrated dc meter. However, for sufficient meter current, the primary would probably need to be excited directly by a signal generator instead of the grid-dip oscillator. The introduction of a signal generator puts this improvement at least into the semiprofessional class. This certainly would be the case if a digital permeability readout were incorporated. Professionalism as such is not frowned upon; things just get more expensive.

A final note is that the permeability of an unknown core can be found even if a coil of relatively few turns is wound on it. Of course, the coil leads are not connected when it is placed in the instrument.

Most of the above effort is concentrated in the hf range; for example, the selsyn-case instrument/grid-dip oscillator dips occurred in the 7-8 MHz range. To investigate the detailed behavior of cores from, say, 30 MHz to vhf many permeability meters would probably have to be built. They would range from the cup size shown here to thimble sizes for vhf and uhf applications.

## acknowledgements

I wish to express my appreciation for the calibration cores of known permeability that were graciously supplied to me by Bill Amidon of Amidon Associates, in North Hollywood, California. Further, without the encouragement and substantial help of Mssrs. Eulalio Vela and H.A. Dickerson, this effort would have been impossible.

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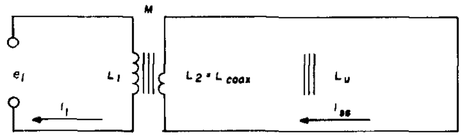
## appendix

### permeability-meter circuit analysis

The following analysis is straightforward ac circuit theory and derives the basic formulas contained in the text. It centers around the three states discussed previously. Subscripts *l* denote the primary coil parameters in all three states. Subscripts *s* and *ss* designate the

secondary winding parameters in states **B** and **C** respectively. Primary- and secondary-loop equations are written for each state as it is discussed. It is assumed that all resistance in the loops is negligible compared with the corresponding inductive reactances so that no resistances will appear in the loop equations. This assumption, besides the simplification in the loop equations it affords, is valid for the inherent high-Q circuitry of the permeability meter itself. When consideration of the unknown toroid dissipation factor is of interest, then these resistances must be included in the analysis. This is done in reference 3.

Consider state **C** (lid on; *i.e.* shorted secondary, and unknown toroid inside) as sketched below. The primary and secondary loop equations are:




STATE C EQUIVALENT CIRCUIT

$e_l = i\omega L_l I_l - i\omega M I_{ss}$ for the primary loop **(1-1)**

$0 = i\omega(L_2 + L_u)I_{ss} - i\omega M I_l$ for the secondary loop **(1-2)**

$e_l$ = applied voltage of the grid-dip oscillator input

$I_l$ and $I_{ss}$ = corresponding loop currents

$M$ = mutual inductance between primary and secondary windings

$L_u$ = unknown toroid inductance;

$L_u = (\mu - 1)\, L_{ua}$, where

$L_{ua} = 0.004046\, h\, log_{10}\, (OD/ID)$ *in* $\mu H$

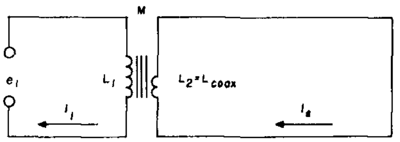
$\omega = 2\pi f$, where $f$ is the grid-dip oscillator input frequency to the primary toroid winding.

Estimating $I_{ss}$ between eqs. **1-1** and **1-2** gives for state **C** input impedance, $Z_{ss}$, and input inductance, $L_{ss}$:

$Z_{ss} = e_l/I_l = i\omega[L_l - M^2/(L_2 + L_u)]$; i.e.,

$$L_{ss} = L_l - [M^2/(L_2 + L_u)] \quad \textbf{(1-3)}$$

For state **B** (lid on, secondary shorted, with no toroidal inside), the loop equations are




STATE B EQUIVALENT CIRCUIT

$e_l = i\omega L_l I_l - i\omega M I_s$ for the primary loop **(1-4)**

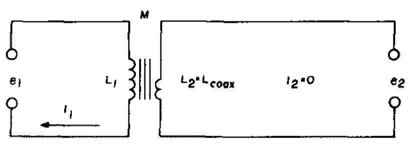
$0 = i\omega L_2 I_s - i\omega M I_l$ for the secondary loop **(1-5)**

where the symbols follow from the above explanation. Elimination of $I_s$ between eqs. **1-4** and **1-5** gives for state **B** input impedance, $Z_s$, and input inductance, $L_s$:

$Z_s = e_l/I_l = i\omega(L_l - M^2/L_2)$; *i.e.,*

$$L_s = L_l - M^2/L_2 \quad \textbf{(1-6)}$$

For state **A** (lid off, secondary open, no toroid inside), where $e_2$ is the open-circuit secondary voltage when grid-dip oscillator voltage $e_l$ is applied, the loop equations are




STATE A EQUIVALENT CIRCUIT

$e_l = i\omega L_l I_l$ for the primary loop **(1-7)**

$e_2 = -i\omega M I_l$ for the (open) secondary, and

$Z_l = e_l/I_l = i\omega L_l$ **(1-8)**

If the leakage inductance is negligible, then dividing **eq. 1-7** by **eq. 1-8** yields

$e_l/e_2 = n = -L_l/M$, where $n$ is the turns ratio, or the number of primary turns for our one turn secondary. **(1-9)**

Substituting into the right-hand portions of **eqs. 1-3** and **1-6**, from **eq. 1-9** to eliminate $M$ and $L_2$ results in (a) an equation for the unknown inductance, $L_u$, namely,

$$L_u = \left[\frac{L_l^2}{n^2(L_l - L_s)}\right]\left[\frac{(L_{ss} - L_s)}{(L_l - L_{ss})}\right] \quad \textbf{(1-10)}$$

or from above, with the unknown inductance usually written in terms of its free-space inductance as $L_u = (u - 1)L_{ua}$, then equivalently,

$$\mu - 1 = \left[\frac{1}{0.004046 h log_{10}(OD/ID)}\right]\left[\frac{L_l^2}{n^2(L_l - L_s)}\right]\left[\frac{(L_{ss} - L_s)}{(L_l - L_{ss})}\right] \quad \textbf{(1-11)}$$

and (b), an equation for the secondary inductance,

$L_2 = L_{coax}$; $L_{coax} = L_l^2/n^2(L_l - L_s)$ **(1-12)**

Notice from **eqs. 1-10** and **1-12** that now $L_u = L_{coax}\,(L_{ss} - L_s)/(L_l - L_{ss})$.

Generally, the grid-dip oscillator resonant dip frequency, $f_i^2$, or $\omega_i^2 = 1/L_iC$, where the inductance, $L_i$ can correspond to $i = 1, 2, s, ss$, and $C$ is the capacitor in the grid-dip oscillator loop. Again, with **eq. 1-10**, using **eq. 1-12**, yields for the unknown inductance, $L_u$ (recalling that $f_s = f_{HI}$, $f_l = f_{LO}$ and $f_{ss} = f_{cor}$ from the text):

$$L_u = (L_{coax} - L_l/n^2)(f_{HI}^2 - f_{cor}^2)/(f_{cor}^2 - f_{LO}^2) \quad \textbf{(1-13)}$$

With $L_u = (\mu - 1)L_{ua}$, from **eq. 1-13** where $f = f_{cor}/f_{HI}$, and $a = f_{LO}/f_{HI}$, gives

$$\mu - 1 = \frac{k}{h log_{10} OD/ID}\,(1 - f^2)/(f^2 - a^2) \quad \textbf{(1-14)}$$

where the instrument constant, $k$, is written as

$k = (L_{coax} - L_l/n^2)/0.004046$ (inductances in $\mu H$) **(1-15)**

$k = (L_{coax} - L_l/n^2)/4.046$, or $k = (L_{coax} - L_l/n^2)/4$ **(1-16)**

where these inductances are now in $nH$.

To obtain **eq. 6** in the text, rewrite **eq. 1-12** as

$$L_{coax} = L_l/n^2 \cdot (1 - L_s/L_l) \quad \textbf{(1-17)}$$

Now $L_s/L_l = f_l^2/f_s^2 = f_{LO}^2/f_{HI}^2$, when using the resonating capacitor, $C$, in the loop. If these frequency ratios are inserted into **eq. 1-17**, where the fractional frequency spread is $\Delta f/f_{LO} = (f_{HI} - f_{LO})f_{LO}$, then

$$L_l/n^2 = \left(1 - \frac{1}{(1 + \Delta f/f_{LO})^2}\right) L_{coax} \quad \textbf{(1-18)}$$

which is **eq. 7** in the text.

To show that the instrument constant, $k$, will change little, as long as the fractional frequency spread, $\Delta f/f_{LO}$, is small compared to one, eliminate $L_l/n^2$ between **eq. 6** and **eq. 1-16** above. This gives $k = L_{coax}/4(1 + \Delta f/f_{LO})^2$. It is therefore seen that $k$ will remain essentially constant for small changes in $\Delta f/f_{LO}$ as long as the latter is small compared to one. Again, this then means that the permeability scale on line 11 of the nomogram of **fig. 4** does not have to be adjusted for small changes in the fractional frequency spread from instrument to instrument.

**Eq. 1-13** for the unknown inductance can be rewritten, using eq. 1-12 and noting that $L_sL_l = a^2$, as

$$L_u = [L_{coax}]\,[a^2(1 - f^2)/(f^2 - a^2)] \quad \textbf{(1-19)}$$

or,

$$L_u + a^2 L_{coax} = L_{coax}a^2(1 - a^2)/(f^2 - a^2) \quad \textbf{(1-20)}$$

which is a hyperbolic relationship between the unknown inductance, $L_u$, and $f^2$, with asymptotes at $f = a$ and $L_u = -a^2L_{coax}$. Since $0<a<f<1$, simply using as a criterion uniform measurement accuracy for the lowest to the highest permeabilities, it is seen that $[L_{coax}]\,[a^2(1 - a^2)]$ should be maximized to make the above hyperbola uniformly as "fat" as possible. Since $L_{coax}$ is fixed by the physical cup dimensions, then for any given $L_{coax}$ it is apparent that the maximum of $[L_{coax}]\,[a^2(1 - a^2)]$ will occur at $a^2 = ½$.

**ham radio**

# how many signals does a receiver see?

While listening on the various ham bands, I am always amazed at signal reports where stations supposedly receive reports of 50 or 60 dB over S9. Since I have always felt that this was an incorrect statement, I went through the difficulty of actual band scanning with a calibrated instrument – in this case a Hewlett-Packard Spectrum Analyzer. A special Rohde & Schwarz omnidirectional active receiving antenna was used. Its design is based upon principles developed by Professor Meinke (Technical University of Munich). With very small mechanical dimensions, this antenna provides the equivalent of 6 dB omnidirectional gain, up to 30 MHz. During a recent 24-hour monitoring period, pictures were taken of the analyzer display approximately every three hours.

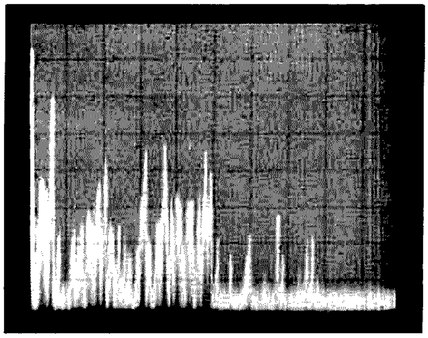
The HP spectrum analyzer was calibrated so that 0 dB on the y-axis was equal to 0 dBm (224 mV) and the scan width was 2 MHz per division, with 20 MHz placed on the extreme right edge of the graticule. The following pictures show the results. The strongest signal (−20 dBm, 22.4 mV into 50 ohms) was the local a-m broadcasting station around 1.6 MHz. This station was located about 10 miles from the receiving site. The most interesting result of this test was that the 15-MHz broadcasting stations were almost as loud as the 80-meter stations, though they were never stronger than −30 dBm. Using the definition that S9 = 50 $\mu$V, these signals were approximately 40 dB over S9. However, no signals of 50 or 60 above S9 could be seen.

These pictures are interesting from another point of view, *i.e.*, they clearly show the change in the maximum usable frequency (MUF). It was most interesting to see that the 15-MHz region was open until very early in the morning and then suddenly the signals disappeared. In addition, the dead zone for the low frequencies can be seen. About midday the propagation on the lower frequencies was extremely poor and then improved later in the day.

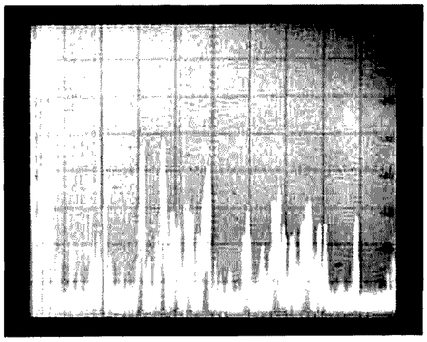
**ham radio**

By **Ulrich L. Rohde, DJ2LR**, 52 Hillcrest Drive, Upper Saddle River, New Jersey 07458

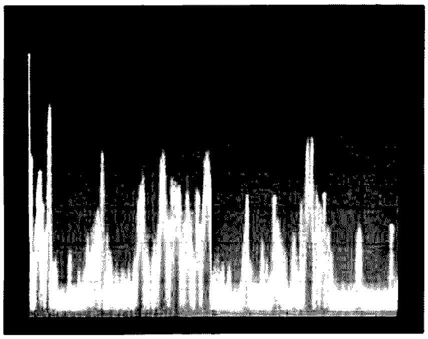
The following pictures were taken from the display of an H-P-141T Spectrum Analyzer system, 8553B rf section and 8552B mainframe. The antenna was a special omnidirectional model that provided 6 dB gain.



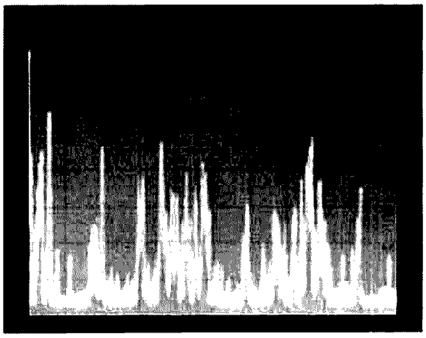
1. 8:00 AM



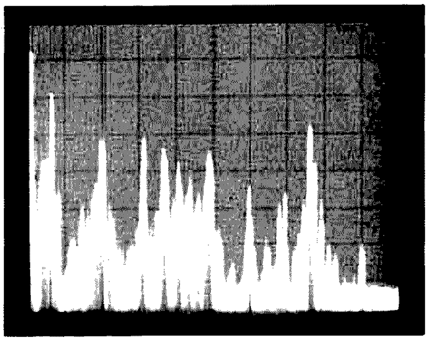
2. 11:30 AM



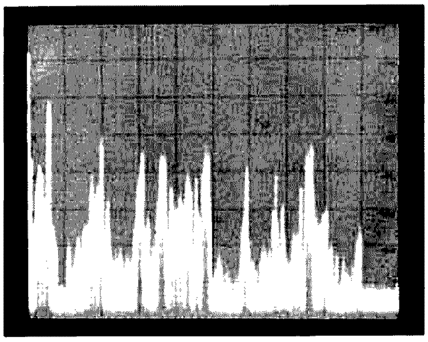
3. 3:00 PM



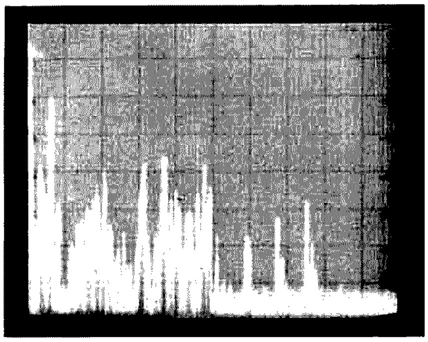
4. 6:00 PM



5. 10:00 PM



6. 1:00 AM



7. 4:00 AM

# short vertical antenna for 7 MHz

## With a good ground system, this antenna will give a good account of itself for those with limited space

Polar radiation patterns in the vertical plane for a low horizontal antenna and a short vertical antenna are shown in **fig. 1**. These patterns are typical for horizontal antennas less than about 0.3 wavelength above the ground and for vertical antennas less than about 0.6 wavelength high. For working beyond a range of a few hundred miles on 7 MHz, only the energy radiated at angles less than about 45 degrees is of much value. **Fig. 1** shows that a low horizontal antenna radiates most of its energy at very high angles, with the maximum being straight up. A low horizontal antenna on 7 MHz is one that is less than about 40 feet (12m) high. Now look at the short vertical pattern. Its radiation straight up is zero and its maximum radiation is at a very low angle, approaching zero. In practice, ground losses change the pattern so that little energy is radiated parallel to the ground. A short vertical antenna for 7 MHz is one that is less than about 80 feet (24m) high.

I erected my antenna in the center of my roof at the peak and ran the radials for the ground plane from that point to the roof edge. Thus, the radials are of different lengths and are not at right angles to the antenna. Instead they slope down. This was the easiest and most logical way to install the radials, and the antenna works well.

All this activity occurred in the fall of the year and I wanted to try the idea quickly, so I made use of the materials I had on hand. For the vertical radiator I used a 10-foot-long (3m) piece of thin-wall electrical conduit with an 8-foot (2.4m) whip attached to the top. I mounted the vertical portion of the antenna on a 5-foot (1.5m) tripod of the type used for television antennas. I made two insulators of plastic to mount the electrical conduit to the tripod. The bottom end of the conduit is about a foot (0.3m) above the roof, and a brass plate with the radials attached is on the roof under the conduit. The two coils of the matching network (discussed later) are mounted to the brass plate with their axes at right angles.

Two things are essential to the success of an antenna of this type. One is an impedance-matching network to allow the efficient transfer of energy from the transmission line to the radiator. The other is a proper ground system.

### antenna impedance

An impedance-matching network is necessary because the impedance of a short vertical antenna is not a good match for most transmission lines. Typically a short vertical antenna will have an impedance consisting of a low resistance and a high capacitive reactance. A conjugate match can be obtained by designing a network to match the transmission line to the resistive part of the antenna impedance and adding an impedance between

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the network and the antenna equal to the antenna reactance but of the opposite sign. Thus for my short vertical, the matching network had to match the transmission line to the few ohms impedance of the antenna. I also included a series inductance having a reactance equal to that of the antenna capacitive reactance. This series inductance compensates for the effect of antenna capacitance.

It would have been desireable to erect the antenna and then measure its impedance at the operating frequency. However, I did not have an impedance bridge. La Port's *Radio Antenna Engineering* contains a good discussion of vertical antennas. A portion of the book is directed to medium-frequency, vertical broadcast antennas, but the data presented seem to work well at the higher frequencies. Curves are presented in the book that show the impedance (resistance and reactance) of vertical antennas as a function of antenna height and antenna length-to-diameter ratio. The diameter of my antenna is not constant since the 8-foot (2.4m) whip is tapered to about 0.125 inch (3mm) diameter at the top. I assumed an average diameter for the entire antenna of 0.5 inch (12.5mm). Thus, the ratio of length to diameter is 18 x 12/0.5 or 432. The antenna height in degrees, where 360 degrees is one wavelength, is 46.3, which is just a little over 1/8 wavelength. Then, from the figures in reference 1, the antenna base resistance will be about 5 ohms and the reactance will be about −j320 ohms. The impedance-matching network thus had to match the 50 ohms of the RG-8/U coax cable I used to 5 ohms. Also, the network had to provide a reactance of +j320 ohms to compensate for the antenna capacitive reactance.

**Table 1** lists the approximate resistance and reactance for several lengths of vertical antennas. This data was obtained from reference 1. In all data shown here, an operating frequency of 7.037 MHz is assumed, but the data should be useful over the entire 40-meter band.

## matching network

I used an L network for the impedance-matching system. These networks are easy to design and work well. I used the equations in reference 2. **Fig. 2** shows an L network. The network can work either way. That is, it can transform from a high to a low impedance (the way

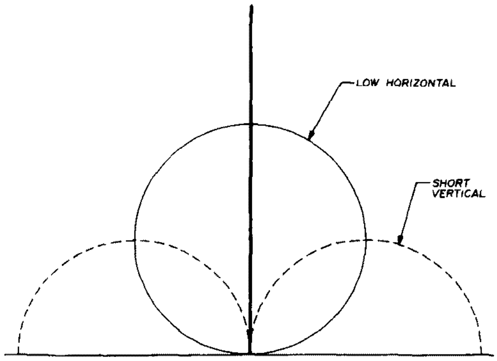


**fig. 1. Typical polar radiation patterns in the vertical plane for a low horizontal and short vertical antenna.**

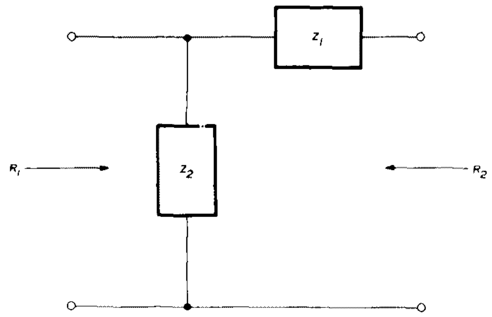


**fig. 2. L network.**

I wanted to go) or the opposite. The network consists of two reactances with opposite signs. That is, one must be capacitive and the other inductive. The values of these reactances are given by:

$$Z1 = \pm j\sqrt{R2(R1 - R2)} \quad (1)$$

$$Z2 = \pm jR1\sqrt{\frac{R2}{R1 - R2}} \quad (2)$$

where
$Z1$ = reactance of the series arm (ohms)
$Z2$ = reactance of the shunt arm (ohms)
$R1$ = resistance (impedance) of the transmission line; 50 ohms for RG-8
$R2$ = radiation resistance of the antenna; 5 ohms in the case of my 18-foot (5.5m) antenna.

In the equations above, $R1$ must always be larger than $R2$. In designing the network, you can use either the two top signs or the two bottom signs of **eqs. 1** and **2**. That is, the series arm can be an inductance and the shunt arm a capacitance, or vice versa. I chose to use an inductance for the shunt arm and a capacitance for the series arm. The reason for doing this is described later.

To match a 50-ohm transmission line to the 5-ohm radiation resistance of the antenna, substitution in **eqs. 1** and **2** gives

$$Z1 = -j\sqrt{R2(R1 - R2)} = -j\sqrt{5(50 - 5)} = -j15 \quad (3)$$

$$Z2 = +j\,R1\sqrt{\frac{R2}{R1 - R2}} = +j50\sqrt{\frac{5}{50 - 5}} = +j16.7 \quad (4)$$

To cancel the effect of the antenna capacitive reactance I added an inductive reactance having the same magnitude, 320 ohms, between the L network and the antenna base.

**Fig. 3** shows the complete matching network. The capacitive reactance of $Z1$ is in series with the added inductive reactance of $+j320$ ohms. Thus the effective series arm is $-j15$ and $+j320$, or $+j305$ ohms. By using the network as shown, the vertical radiator is connected electrically to the ground plane. Thus, if the ground plane is connected to an earth ground the entire antenna system can be grounded.

The value of the two inductances in the complete matching network can be calculated from

$$X_L = 2\pi f L \qquad (5)$$

where

$X_L$ = reactance of the inductance (ohms)
$f$ = frequency of operation (Hz)
$L$ = inductance (H)

The known values of $X_L$ and $f$ are substituted into **eq. 5** and the equation then is solved for $L$. For the shunt arm, having a value of 16.7 ohms and a frequency of 7.037 MHz or $7.037 \times 10^6$ Hz, **eq. 5** becomes

$$16.7 = 2\pi \cdot 7.037 \cdot 10^6 \cdot L1 \qquad (6)$$

and

$$L1 = 0.378 \cdot 10^{-6}\ H = 0.378\ \mu H \qquad (7)$$

Similarly for the series arm, having a value of 305 ohms,

$$305 = 2\pi \cdot 7.037 \cdot 10^6 \cdot L2 \qquad (8)$$

and

$$L2 = 6.90 \cdot 10^{-6}\ H = 6.90\ \mu H \qquad (9)$$

The number of turns required for the two inductances can be calculated if coil length and diameter are assumed. A formula given in the *ARRL Handbook* is:

$$L = \frac{a^2 n^2}{9a + 10b} \qquad (10)$$

where
$L$ = inductance ($\mu$H)
$a$ = coil radius (in.)*
$b$ = coil length (in.)*
$n$ = number of turns in coil

Rearranging **eq. 10**,

$$n = \left[\frac{L(9a + 10b)}{a}\right]^{1/2}$$

The shunt coil is 0.378 $\mu$H and requires 2.7 turns if the diameter and length are 3 inches (76.5mm). Similarly, for the series coil (6.9 $\mu$H), 18 turns are required if the diameter is 3 inches (76.5mm) and the length is 9 inches (230mm).

Even with modest power the current in these coils can be fairly large, so the coils should be constructed accordingly. I used 1/4-inch (6.5mm) copper tubing for the shunt coil and 1/8-inch (3.0mm) copper tubing for the series coil. The coils should be mounted at right angles to each other to minimize inductive coupling.

## ground system

The ground system used with a short vertical antenna is a big factor in the performance of the antenna. The ground system consists of a number of radials, which are wires that extend out from the base of the antenna and in a direction more or less normal to the antenna. Imagine a wheel laid flat on the ground or on a roof with the axle sticking straight up and the wheel spokes extending outward from the axle. The axle represents the radiating part of the antenna and the spokes represent the radials.

* If a and b are in mm, the factors 9 and 10 in the denominator of **eq. 10** are 229.5 and 255 respectively. **Editor**

**table 1. Approximate resistance and reactance for several lengths of vertical antenna at 7 MHz.**

| length feet | (m) | diameter inches | (mm) | length/diameter | length degrees | resistance (ohms) | reactance (ohms) |
|---|---|---|---|---|---|---|---|
| 15 | (4.6) | 0.5 | (12.5) | 360 | 38.6 | 4.5 | −j450 |
| 15 | (4.6) | 0.75 | (19) | 240 | 38.6 | 3.5 | −j400 |
| 15 | (4.6) | 1.0 | (25.5) | 180 | 38.6 | 2.9 | −j380 |
| 18 | (5.5) | 0.5 | (12.5) | 432 | 46.3 | 5 | −j320 |
| 18 | (5.5) | 0.75 | (19) | 288 | 46.3 | 4 | −j300 |
| 18 | (5.5) | 1.0 | (25.5) | 217 | 46.3 | 3.5 | −j280 |
| 20 | (6) | 0.5 | (12.5) | 480 | 51.3 | 8 | −j270 |
| 20 | (6) | 0.75 | (19) | 320 | 51.5 | 5 | −j250 |
| 20 | (6) | 1.0 | (25.5) | 240 | 51.5 | 4.5 | −j230 |
| 20 | (6) | 1.25 | (32) | 192 | 51.5 | 4 | −j200 |
| 25 | (7.6) | 0.5 | (12.5) | 600 | 64.3 | 14 | −j160 |
| 25 | (7.6) | 0.75 | (19) | 400 | 64.3 | 12 | −j150 |
| 25 | (7.6) | 1.0 | (25.5) | 300 | 64.3 | 11 | −j140 |
| 25 | (7.6) | 1.25 | (32) | 240 | 64.3 | 10 | −j130 |
| 30 | (9) | 0.5 | (12.5) | 720 | 77.2 | 22 | −j55 |
| 30 | (9) | 0.75 | (19) | 480 | 77.2 | 22 | −j50 |
| 30 | (9) | 1.0 | (25.5) | 360 | 77.2 | 22 | −j45 |
| 30 | (9) | 1.25 | (32) | 288 | 77.2 | 22 | −j41 |
| 30 | (9) | 1.5 | (38) | 240 | 77.2 | 22 | −j40 |

Most a-m broadcast stations use vertical antennas and a ground system consisting of 120 radial wires spaced every 3 degrees from the base of the antenna. Radial length seldom exceeds a half wavelength and often is less.

Reference 1 shows the field strength produced by a vertical radiator as a function of the height of the antenna and the number and length of the radials in the ground system. Two conclusions are apparent from the information presented: (1) use as many radials as you possibly can, and (2) make the radials as long as possible. With regard to the number of radials, broadcast stations use 120, which gives an efficiency close to the theoretical maximum. There probably are not too many amateurs who will use 120 radials, but use as many as you possibly can.

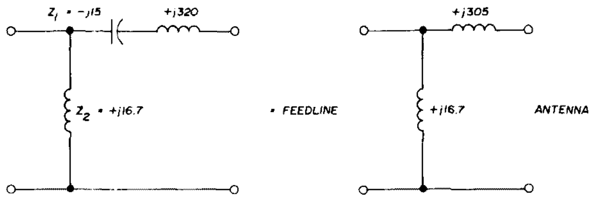


**fig. 3. Complete matching network for the short vertical antenna.**

With regard to the length of the radials, there is little to be gained by making the length greater than a half wavelength. For many it's not possible to even approach a half wavelength for the radials; there simply is not that much room on a city lot or on the average rooftop. If you can't make the radials that long, you can compensate somewhat for their shortness by using more of them. Thus again, the rule is to use as many radials as you possibly can and make them as long as possible. The shorter the antenna, the more important this is.

The ground system for a broadcast antenna is usually

buried a few inches to get it out of the way, and the wires are at right angles to the vertical radiator. With my roof-mounted vertical antenna, the radials are on the roof and slope away from the vertical radiator instead of being at right angles. However, this antenna performs well, so the sloping radials apparently do not effect performance adversely.

In my system I mounted an 8 x 10 inch (20x25cm) brass plate under the lower end of the vertical radiator. I drilled a series of holes around the edge of this plate and tapped them for 6-32 (M3.5) machine screws. When I installed the radial wires I attached each one under the head of a screw mounted in one of the tapped holes. After I had attached all of the radials to the plate I soldered the wires and the screws to the base plate.

My house is single story about 29 x 41 feet (9x12m) with a hip roof. I mounted the antenna in the center of the roof at the peak. I used 20 radials of no. 24 (0.5mm) bare copper wire and spaced them equally. My longest radial is about 25 feet (7.6m); the shortest is about 15 feet (4.6m).

## adjustment

Adjustment of the matching-network series coil is necessary. As explained earlier, the purpose of this coil is to compensate for the antenna capacitive reactance. The information about the capacitive reactance of the antenna is approximate rather than exact. Thus it's necessary to either measure the reactance of both the antenna and series coil and make them equal or adjust the inductance after it is in place and make it "tune out" the antenna capacitance. I did the latter.

With the antenna, ground system and feed line installed, I used a vswr meter at the transmitter and adjusted the series-coil inductance for the lowest swr at the low end of the 7-MHz band. I obtained a swr of 1.5:1. I adjusted the coil by spreading or compressing the length of the coil. This adjustment is critical. It should be possible to perform this adjustment with a noise bridge or with a rf-current indicator connected at the junction of the series coil and the vertical-antenna base. For the noise bridge, adjust the series coil so the antenna is resonant at the desired frequency. For the rf-current meter, adjust the coil for maximum current.

## results

This short vertical antenna has given a good account of itself. If your antenna farm is smaller than you'd like, or if you can't erect a 70 footer (20m) consider a vertical, even a short one. Consider also installing the antenna on top of the house. Installing the antenna on the roof reduces clutter in the back yard and allows the antenna to "see" the horizon much better than at ground level.

## references


1. E. A. La Port, *Radio Antenna Engineering*, 1952 edition, McGraw-Hill Book Company, New York.
2. F. E. Terman, *Electronic and Radio Engineering*, 4th edition, McGraw-Hill Book Company, New York, page 115.


**ham radio**

# video modulated four-tube amplifier for 1270-MHz television

## Construction details of a four-tube ATV amplifier which provides continuous-power output of 120 watts at 1270 MHz

The **need to develop adequate** power on fast-scan television, at 1270 MHz, came about in 1974 when the Chicago area ATV group decided to look to bands other than 420-450 MHz for A5 operation. Results with low power and small antennas proved to be very useful and it was felt that higher power and better antennas would help to eliminate some problems with snow and improve the range.

After consulting with George, W9WCD, and reading the *ham radio* article by WB6IOM, I decided to try a cavity amplifier using four 2C39A tubes in a similar ring cavity. The only requirement not covered by the WB6IOM amplifier was the plate-circuit tuning range. I decided that it would be desirable to tune the entire 1215 to 1300 MHz range because of the great deal of radar interference. This tuning range would also allow us to stay away from any interference resulting from harmonics of 439 and 427 MHz ATV transmissions. It was found that the radar and other undesired transmissions were at a minimum at 1266 MHz and full ATV duplex was very easily accomplished.

### 1270 MHz amplifier

The complete four-tube amplifier is mounted on a 17x13x3-inch (43.2x33.0x7.6cm) chassis, complete with modulator, modulator power supply, air cooling system, video detector, and filament transformer. The amplifier is operated with the grids at dc and rf ground; bias is supplied from the direct-coupled modulator. For normal video-modulated operation, drive requirements are about 15 watts, at a plate current of 400 mA and a plate

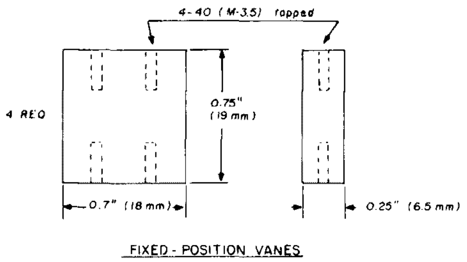


**fig. 1. Anode cavity fixed-position vanes.**

voltage of 800 volts. Normal power output from the amplifier is about 120 watts average, as measured on a Bird model 43 wattmeter. Since the amplifier is used for continuous operation, it is suggested that adequate cooling and moderate plate voltage and current be used. It is also important that the power supplies are rated for continuous service.

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The amplifier was built with the aid of a lathe and drill press. It is built entirely of brass and bronze that was obtained mainly from the local junk yards. The circular brass flat material was initially rough cut from 0.08-inch (20mm) sheet stock using a power sabre saw and then cut to size on the lathe with the use of a face

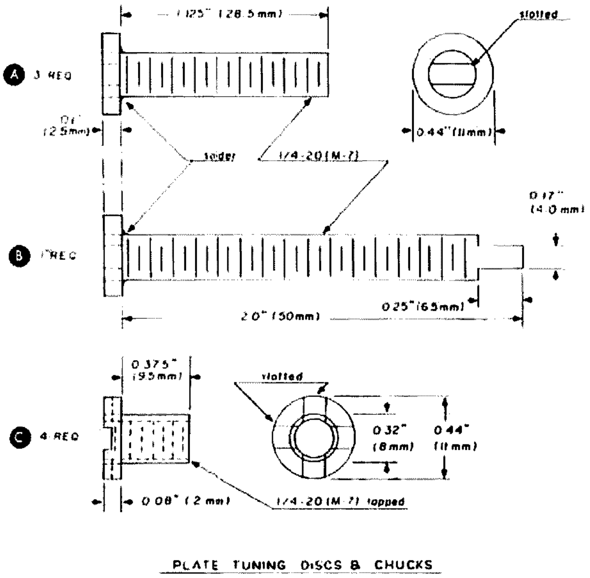



fig. 2. Plate-tuning disks and chucks. The three short tuning disks are slotted to permit adjustment during initial tune-up of the amplifier.

plate. The large clearance hole on the circular disks was also cut on the lathe.

## anode cavity

The four 2C39As are placed around a 2.57 inch (6.5mm) diameter circle within a cavity of 4.741 inch (12cm) OD and a height of 0.74 inch (1.9cm). The natural resonant frequency of the cavity is below the 1270-MHz band but is raised by the use of four 0.69x0.74x0.25-inch (17.5x19.0x6.5mm) fixed-position

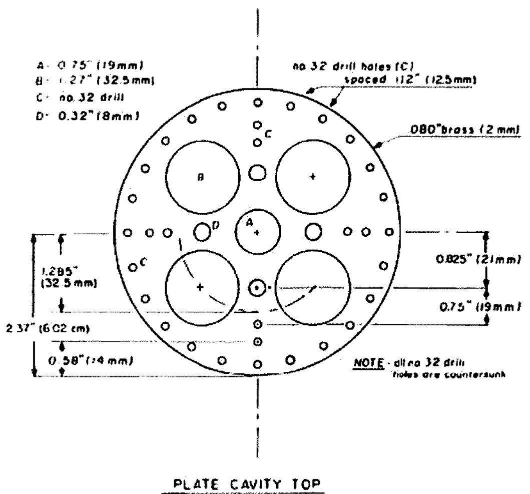
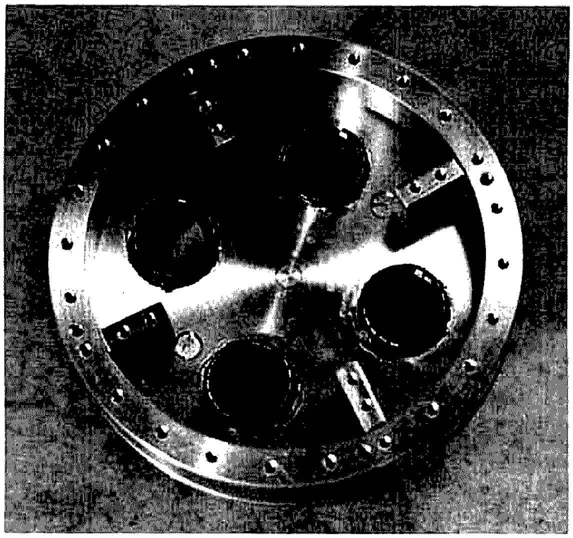



fig. 3. Top plate for the anode cavity.



Looking into the anode cavity from the top. The four fixed-position vanes are installed.

brass vanes (fig. 1). The vanes are place inside and bolted to the top and bottom walls through countersunk, tapped 4-40 (M3.5) holes.

Four plate-tuning disks, placed in a 1.65-inch (4.2cm) diameter circle, and centered between each tube, adjust the plate-circuit resonant frequency. Only one tuning disk is required for external plate-tuning adjustment. The others are fixed and placed in approximately the same position within the cavity. The four plate-tuning disks (see fig. 2) are machined from brass stock and ¼-20 M7 threaded brass rod. The 0.44-inch (1.1cm) disk is drilled in the center, large enough to force fit a small filed-down portion of the threaded rod. The rod is then soft soldered to the rear of the disk.

A 0.357x0.44-inch (9.5x11.0mm) chuck is machined from brass stock and then drilled and tapped in the center for the ¼-20 (M7) shank of the plate tuning disks (fig. 2C). A fine-bladed saw is used to slot the walls of the chuck so they can be compressed and adjusted for a snug fit. Four holes (0.32 inch, [8mm]) are drilled in the anode cavity top plate (fig. 3) and each chuck is inserted and soft soldered in place with the aid of a propane torch.

The output-coupling probe (fig. 4) is built by fashioning machined brass tubing, rod, flat stock, and Teflon rod into a length of concentric line (coax). The shield is made by inserting and soldering the ½-inch (12.5mm) OD brass tubing into the cable end of a UG-23B/U type-N connector; the end of the connector must be enlarged to accept the tubing. The center conductor of the probe is made from a small piece of 0.123-inch (3mm) diameter brass rod. One end is attached to the center pin of the connector while the other is cut for 6-32 (M3.5) threads. The Teflon rod is machined to fit inside the brass tubing. A hole is drilled through the center of the Teflon rod to accept the brass rod. The probe's disk (0.48 diameter, x 0.1 inch thick [12x2.5mm]) is drilled and tapped for 6-32 (M3.5)

threads. The disc is then screwed onto the probe assembly and held by a stop nut. A 0.975x0.75-inch (25x19mm) chuck (fig. 5) for the output-coupling probe was machined from brass stock and slotted for a snug fit to the output-coupling probe. The chuck is also soldered to the underside of the center hole in the top plate of the anode cavity.

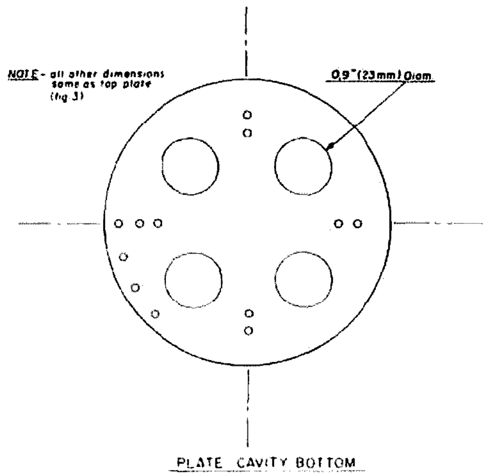
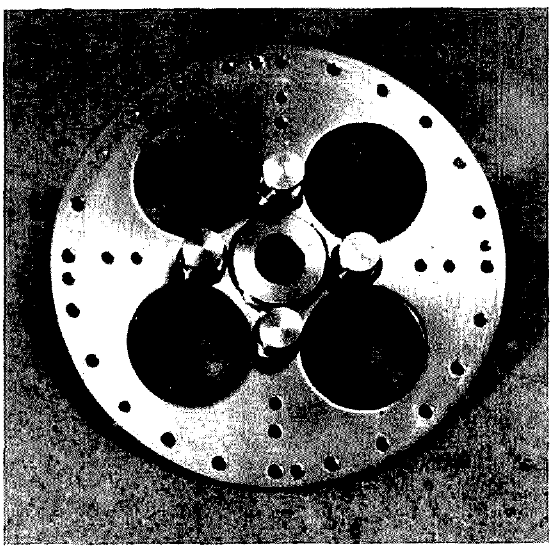



**fig. 7. Bottom plate for the anode cavity.**

The plate connector and bypass disk (fig. 6) is 0.08-inch (20mm) thick by 4.742-inch (12cm) OD with appropriate clearance holes for the tubes, plate-tuning chuck, and output-coupling chuck. The finger stock for the anode connector (Instrument Specialty Company 97-70) is soldered in the four tube holes on the disk. A piece of 0.01-inch (0.5mm) Teflon sheet stock is used as the insulation for the plate-bypass capacitor. Clearance holes are cut out for the four tubes and the chucks.

Homemade Teflon shoulder insulators were used to



**Anode cavity top plate showing the four tuning disks. The holes have been countersunk on the other side to permit the Teflon insulator to lie flush against the plate.**

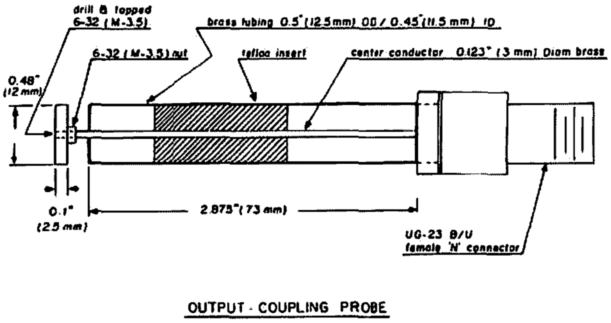



**fig. 4. Output-coupling probe.**

mount the plate connector and bypass disk to the top of the plate cavity. The anode cavity top and bottom plates (fig. 7) are bolted to the side wall portion by 4-40 (M3) screws. The countersunk holes are spaced every ½ inch (12.5mm) on the top plate and every 5/8 inch (16mm)

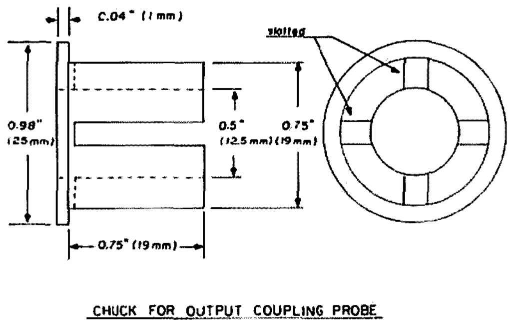



**fig. 5. Chuck for the output-coupling probe.**

on the bottom plate. The bottom plate contains the finger stock (Instrument Specialty Company 97-74) for grounding the grid connections.

The cathode compartment shown in fig. 8, measures 3.98 inch (10.1cm) OD by 2.05 inch (5.2cm) deep and

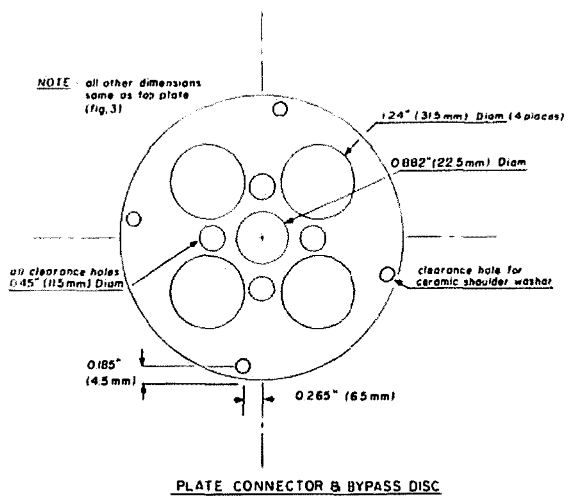
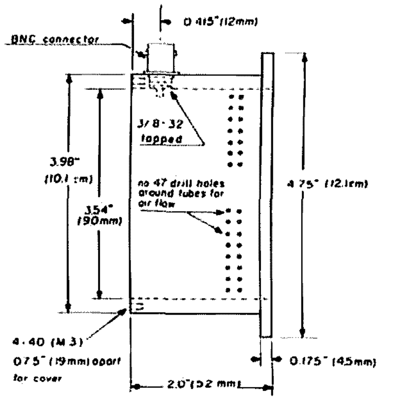



**fig. 6. Plate connector and bypass disk. Each tube hole has Instrument Specialty Company 97-70 finger stock soldered on the inside. The Teflon sheet is also cut to the same dimensions.**




CATHODE COMPARTMENT (SIDE)

**fig. 8. Side view of the cathode compartment. The ring that bolts to the anode cavity is shown in its normal position.**

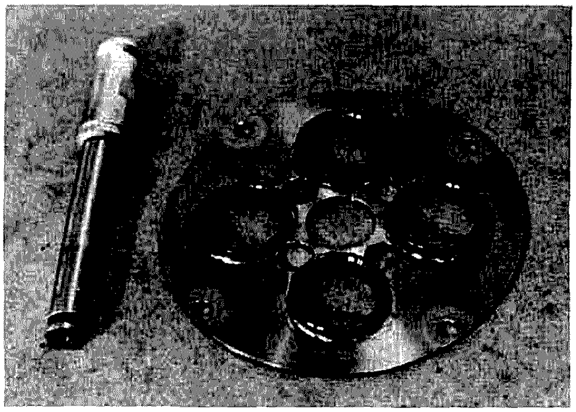
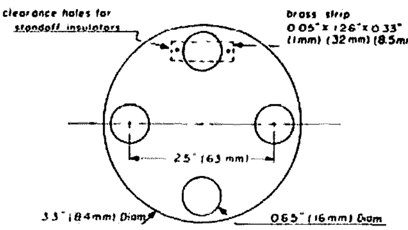
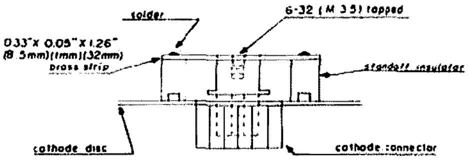


**Plate bypass and connector disk shown with the output-coupling probe.**



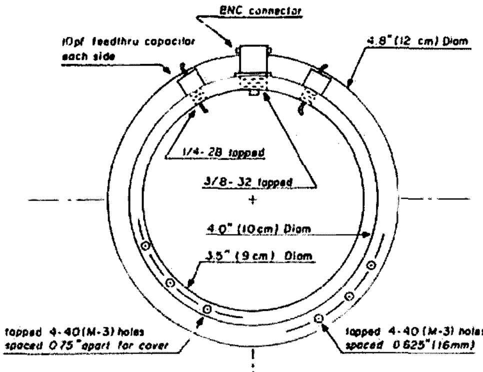

CATHODE DISK

**fig. 9. Cathode disk.**




FILAMENT & CATHODE CONNECTORS

**fig. 10. Filament and cathode connectors.**

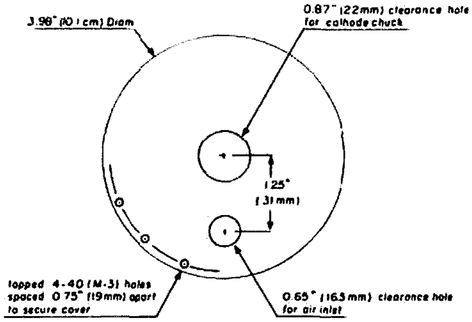



END-VIEW CATHODE COMPARTMENT

**fig. 11. End view of the cathode compartment showing the placement of the video input connector and the filament feed through capacitors.**

contains the cathode and filament connections for the 2C39 tubes. Its side wall is soldered to a 4.755-inch (12.1cm) OD by 3.98-inch (10.1cm) by 0.175-inch (4.5mm) brass ring. This allows the cathode compartment to be bolted to the anode cavity assembly by 4-40 (M3) screws spaced 5/8 inches (16mm) apart.

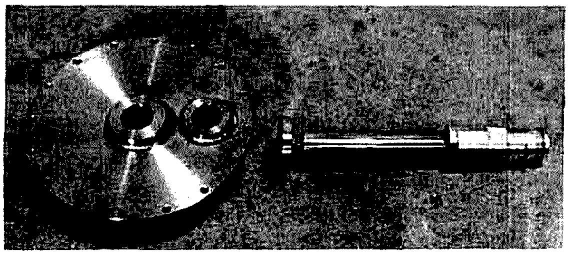
The cathode disk is a piece of 0.08-inch (20mm) thick brass cut into a 3.3 inch (8.4cm) diameter circle (fig. 9) with four 0.65-inch (16.5mm) holes in which Instrument Specialty Company 97-76 finger stock is fitted and soldered. The four filament connectors, shown in fig. 10, are made with two 3/8-inch (9.5mm) high insulated standoff terminals supporting a 1.26-inch (3.2cm) by 0.33-inch (8.5mm) by 0.05-inch (1mm) brass strip. The filament connectors, Instrument Specialty Company 97-280, are centered in the middle of the strips.




CATHODE COMPARTMENT COVER

**fig. 12. Cathode compartment cover.**

A UG-625B/U BNC connector is screwed into the tapped hole into the cathode compartment outside wall. The center pin is connected to the cathode disk through a 10-turn rf choke. As shown in (fig. 11) two 10 pF feed through capacitors are also screwed into the tapped holes. They are used to furnish filament voltage to the tubes. The cathode circuit used in this amplifier is similar to the circuit used in the WB6IOM amplifier.



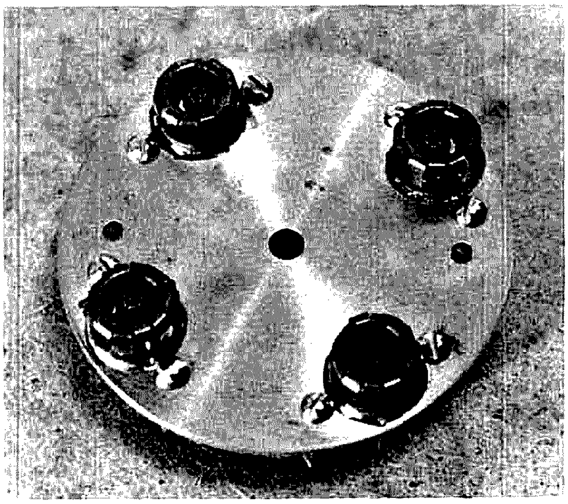
**Cathode compartment cover disk with the input probe.**

A 0.08-inch (20mm) by 3.98-inch (10.1cm) diameter disk with a 0.87-inch (22mm) center hole is used as the cathode compartment cover (fig. 12). A clearance hole (0.65 inches [16.5mm]) is also provided in the cover for an 1/8-inch (3mm) long air-inlet hose connector (fig. 13) which is soldered in place.

The cathode probe is 2.25 inches (5.7cm) long with another type-N connector on one end and a 0.978-inch (26.5mm) disk on the other. The assembly shown in fig. 14 is assembled in the same manner as the anode probe. The chuck for the input probe is machined to an 0.87-inch (22mm) OD, 1.125 inches (28.5mm) long (fig. 15) and is soldered in place after being inserted into the cathode compartment cover center hole. The chuck is slotted and adjusted for a snug fit to the input probe.

## cooling

Cooling for the amplifier is provided by a Dayton (model 2C781) 100-cfm, squirrel-cage, blower mounted

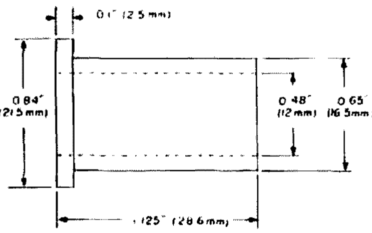


**Completed cathode disk showing the filament and cathode connections.**

on the rear of the main chassis. The air is forced into the enclosed chassis and then directed at the anode cooling fins by a plexiglass manifold and a cutout in the chassis. Air for the filament and grid seals is provided by a 5/8 inch (16mm) ID hose inserted into a chassis connector, similar to the one mounted on the amplifier, and the hose connector on the cathode compartment cover.

## tuning up

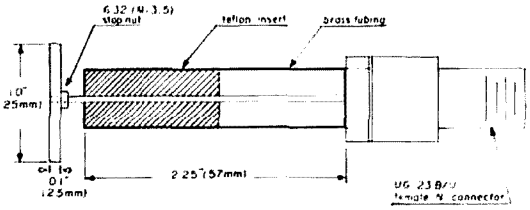
It is suggested that before power is applied to the amplifier a few accessories be available. A 50-ohm



AIR INLET HOSE CONNECTOR

**fig. 13. Air inlet hose connector.**

dummy load, or resonant antenna, and a power indicator is necessary for proper amplifier tune-up. To prevent destroying the modular output transistors, the output cable is removed and an adjustable wirebound, 100-ohm 25-watt potentiometer is used to replace the modulator.

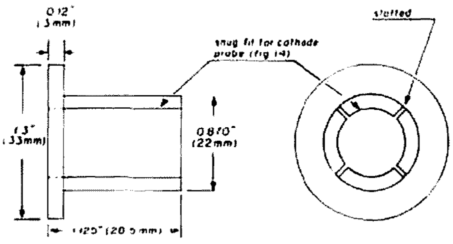



NOTE: refer to fig 4 for relevant information

INPUT - CONNECTOR PROBE

**fig. 14. Cathode input probe.**

Use 2C39A tubes that you know to be good. With proper cooling and about 500 volts applied to the amplifier, adjust the 100-ohm cathode resistor for a plate current of 100 mA. Now apply about 5 watts of drive power and adjust the cathode probe for maximum plate current. Now resonate the amplifier plate circuit. This is




CHUCK FOR CATHODE PROBE

**fig. 15. Cathode probe chuck.**

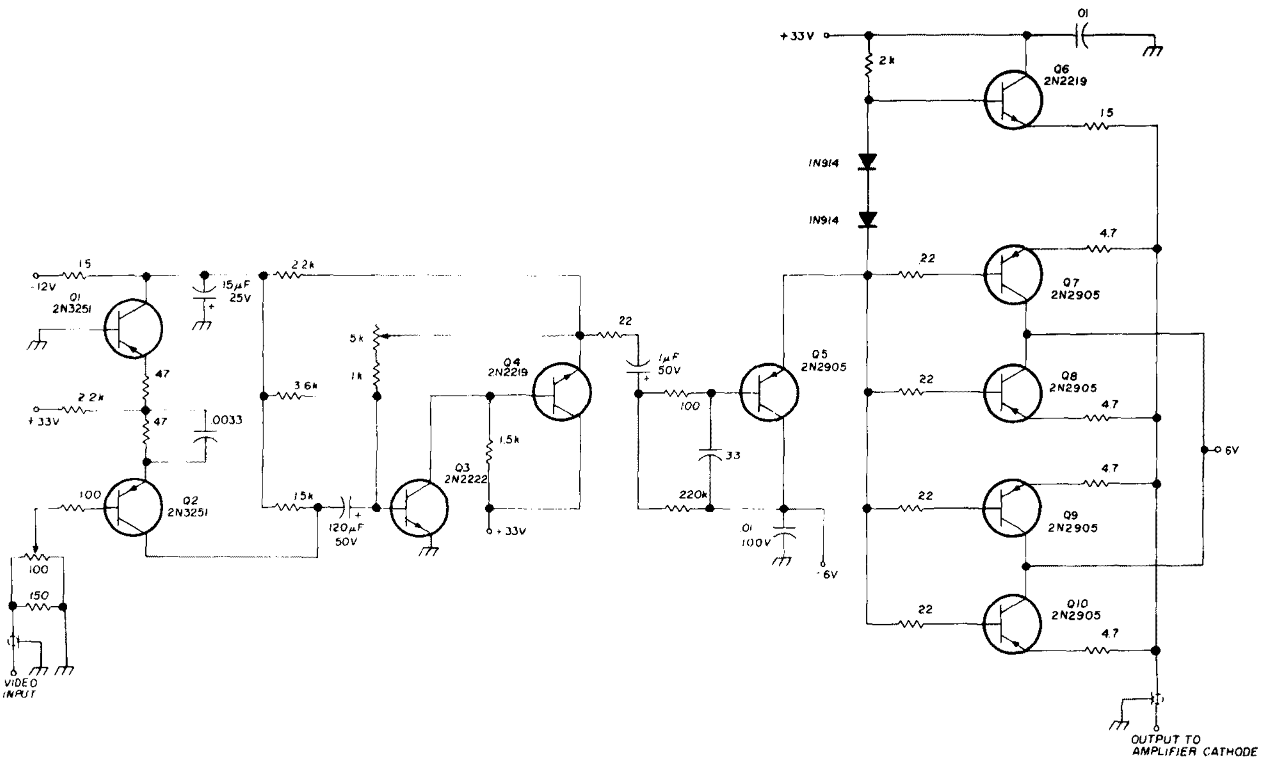


fig. 16. Schematic diagram of the cathode-video modulator. All resistors are ¼-watt. The supply voltages go to the modulator through 0.001 μF feedthrough capacitors.

done by starting with the four plate-tuning disks at about midrange. While closely observing the power-output indicator, vary the four tuning adjustments several turns in or out until there is power output. Since the output-probe adjustment and plate-tuning adjustments interact it's necessary to repeat all adjustments

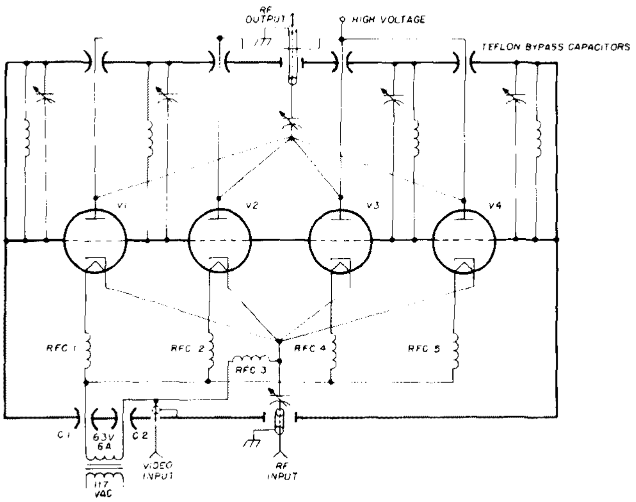


fig. 17. Schematic representation of the four-tube 1270-MHz amplifier. C1 and C2 are 10 pF, RFC1 through RFC5 are each 10 turns no. 22 AWG (0.6mm) airwound on 1/8-inch (3mm) diameter forms.

until maximum output power is obtained. All four tuning adjustments should be positioned equally and the output-coupling probe position optimized before the plate voltage is increased. After the plate voltage has been raised to 800 volts, the tuning must be readjusted for maximum output.

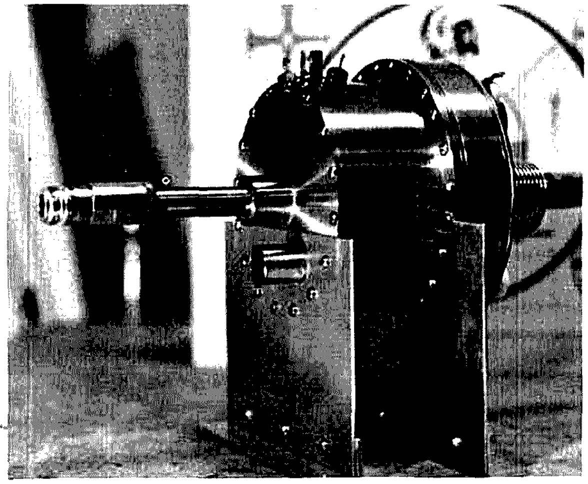
The video modulator may now be reconnected to the amplifier, and with the camera connected to the modulator it should be possible to obtain excellent pictures from an inline detector on the output transmission line.

## cathode video modulator

The modulator circuit used for the amplifier is very similar to one in use by a uhf television station in the Chicago area. However, higher power-supply voltages were used to increase the video output swing, and emitter peaking was used on Q2. I also found it was desirable to include a bias adjustment of Q4 and to include two more 2N2905 transistors in the output stage to handle the additional cathode current. Originally this circuit was used to modulate a single 2C39A which was operated at lower plate current than normal. Normal video output level from closed-circuit TV cameras is 2 volts p-p ±1.5 volts, which is adequate for the modulator.

The power-supply requirements are −12 volts at 35 mA, +33 volts at 70 mA, and +6 volts at 600 mA. All voltages are regulated by 10-watt Zener diodes. Transistor Q5 is cooled with a clip-on heatsink.

The modulator is built on a piece of copper-clad

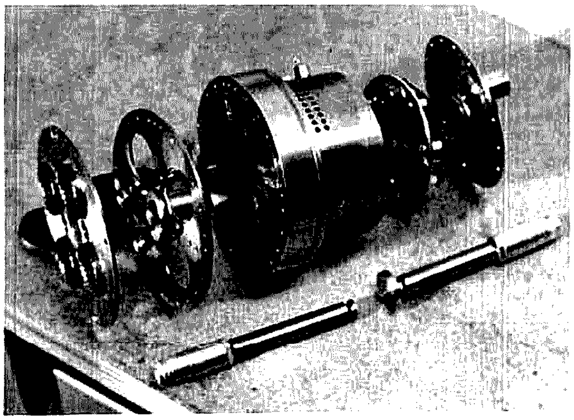


Rear view of the completed amplifier. Note the sandwich-type construction of the individual plates. Starting from the left: cathode comparment cover plate, cathode compartment, anode cavity bottom plate, anode cavity, anode cavity top plate, Teflon insulator, and anode bypass and connector plate.

printed-circuit board, mounted in a 2x3½x6¼-inch (5x8.9x15.9cm) Minibox. Stages Q7 through Q10 are mounted on an insulated 1-3/4x4-3/4 inch (4.5x12cm) heatsink on the top of the Minibox. The modulator output is directly coupled to the four-tube cavity through a short length of RG-59/U coaxial cable.

## acknowledgements

I would like to thank the people in the Chicago area who worked together to prove that excellent results on ATV may be accomplished on the 1270-MHz band.



Disassembled amplifier showing the individual components.

Stations that participated in the venture include WA9CGZ, W9LK, W9DUT, WA9EUN, K9HDE, W9NAU, and W9YTM. It may be note worthy that with the many megahertz available in this band, it seems wasteful to tie up a large portion of the 440-MHz band with ATV repeaters.

**ham radio**

MICROPROCESSORS

# microprocessor interfacing: register pair instruction

**In our last installment** we discussed the single-byte data transfer instructions, MOV D, S and MVI <B2>, in the 8080 microprocessor instruction set. **Fig. 1** summarizes the significant points by indicating that for MVI r <B2> instruction, the data byte transferred to a register comes from the instruction itself, whereas for the MOV D, S instruction, the data byte is copied into the destination register D from the source register S.

Since the 8080 chip has a 16-bit address bus, there exists within the chip both a 16-bit *program counter* and a 16-bit *stack pointer*.[1] Due to this architecture, it would be very convenient to be able to manipulate full 16-bit address words within the chip. This is achieved by using the *register pair* operations. The six general purpose registers are treated as three sets of register pairs, B and C, D and E, and H and L. Each register pair is then designated by a unique 2-bit register pair code,

| register pair | HI byte | LO byte | 2-bit register pair code |
|---|---|---|---|
| B | B | C | 00 |
| D | D | E | 01 |
| H | H | L | 10 |

The final 2-bit register pair code, 11, is reserved for either the stack pointer (SP) or the program status word (PSW), which contains the contents of the accumulator and the flag bits. *In register pair operations*, the HI byte is always the most significant eight bits in the 16-bit memory address. The LO byte is the least significant eight bits. Registers B, D, and H function as HI address bytes, and C, E, and L as LO address bytes.

As one example of a register pair operation, consider the three-byte *load register pair immediate* instructions:

| | |
|---|---|
| LXI rp | |
| <B2> | [LO byte] |
| <B3> | [HI byte] |

This permits you to move the data bytes contained within the second and third bytes of the instruction to the register pair, rp. The general form of the instruction is

| 0 0 | R P O | 0 0 1 |
|---|---|---|
| Instruction class | 2-bit code for register pair | |

Some examples include:

| data transfer operation | mnemonic | octal instruction code |
|---|---|---|
| <B2> → C | LXI B | 001 |
| | <B2> | <B2> |
| <B3> → B | <B3> | <B3> |
| <B2> → E | LXI D | 021 |
| | <B2> | <B2> |
| <B3> → D | <B3> | <B3> |
| <B2> → L | LXI H | 041 |
| | <B2> | <B2> |
| <B3> → H | <B3> | <B3> |

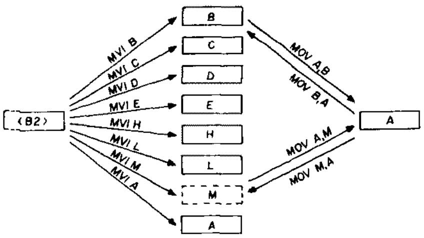



**fig. 1. Schematic representation of the MVI r and several MOV D,S instructions.**

It should be noted that the accumulator and "memory register" are not used as a register pair. To transfer data from memory location HI = 030 and LO = 123, you execute the following simplified program:

| octal instruction code | mnemonic | comments |
|---|---|---|
| 041 | LXI H | Load register pair H with the following LO and HI address bytes |
| 123 | 123 | L register byte |
| 030 | 030 | H register byte |
| 126 | MOV D, M | Move data from the memory location addressed by register pair H to register D |

**By Peter R. Rony, Jonathan Titus, and David G. Larsen, WB4HYJ**

Mr. Larsen, Department of Chemistry, and Dr. Rony, Department of Chemical Engineering, are with the Virginia Polytechnic Institute and State University, Blacksburg, Virginia. Mr. Jonathan Titus is President of Tychon, Inc., Blacksburg, Virginia.

Fig. 2 summarizes the four LXI rp instructions. Keep in mind that you cannot arbitrarily pair registers. For example, if you wish to load registers C and D with data for an operation, you will not be able to use an LXI rp instruction; use MVI C and MVI D instead. If you could substitute register E for register C, then you can use an LXI D instruction to load both register bytes into the indicated registers. Other useful register pair operations shown in fig. 2 are:

| register pair operation | octal instruction code | comments |
|---|---|---|
| XCHG | 353 | Exchange the contents of register pair H with the contents of register D. The HI bytes, H and D, exchange with each other and the LO bytes, L and E, exchange with each other. |
| SPHL | 371 | Load the contents of register pair H into the stack pointer. |
| PCHL | 351 | Load the contents of register pair H into the program counter. |
| LXI SP<br><B2><br><B3> | 061<br><B2><br><B3> | Move instruction bytes <B2> and <B3> into the stack pointer. <B2> is the LO byte and <B3> is the HI byte. |

Since the program counter always contains the address of the next instruction to be executed, the register pair instruction PCHL is actually a jump instruction.

Two other useful instructions for manipulating 16-bit memory address words are the *increment register pair*, INX rp, and *decrement register pair*, DCX rp, instructions.

| 0 0 | R P 0 | 0 1 1 |
|---|---|---|
| instruction class | 2-bit code for register pair | increment operation |

| 0 0 | R P 1 | 0 1 1 |
|---|---|---|
| instruction class | 2-bit code for register pair | decrement operation |

With these instructions, the 16-bit contents of the register pair are either incremented or decremented as a single 16-bit word. However, no condition flags are affected by the INX rp or DCX rp instructions. You do not get a carry out of the most significant bit, the zero flag is not set when the register pair contents are zero, etc. It should be clear that the INX rp and DCX rp instructions are designed for 16-bit address operations, not for multiprecision arithmetic.



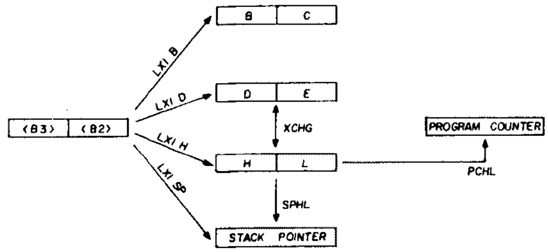



fig. 2. Representation of several register pair instructions, including LXI rp, XCHG, SPHL, and PCHL.

Other register pair operations include:

| register pair operation | octal instruction code | comments |
|---|---|---|
| XTHL | 343 | Exchange the top of the stack with the contents of register pair H |
| DAD rp | 011, 031, 051, or 071 | Add contents of register pair rp to register pair H. Only the carry flag is affected. |
| PUSH rp | 305, 325, 345, or 365 | Push register pair rp on stack |
| POP rp | 301, 321, 341, or 361 | Pop register pair rp off stack |
| SHLD<br><B2><br><B3> | 042<br><B2><br><B3> | Move contents of register L to memory location specified in instruction bytes <B2> and <B3>. Move contents of the H register to the succeeding memory location. |
| LHLD<br><B2><br><B3> | 052<br><B2><br><B3> | Load register L with contents of memory location specified in instruction bytes <B2> and <B3>. Load register H with the contents of the succeeding memory location |

The SHLD and LHLD instructions can be used as 16-bit I/O instructions if the input-output devices are addressed as memory locations.[2,3]

These instructions will be discussed in subsequent columns.

## references

1. P. R. Rony, J. A. Titus, and D. G. Larsen, "Microcomputer interfacing: Internal registers within the 8080 chip," *ham radio*, February, 1977, page 63.
2. P. R. Rony, J. A. Titus, and D. G. Larsen, "Microcomputer interfacing: Accumulator I/O *vs* memory I/O," *ham radio*, June, 1976, page 64.
3. B. Gladstone and P. D. Page, "Programming Hints Ease Use of Familiar Microprocessor," *Computer Design 15*, August, 1976, page 77.

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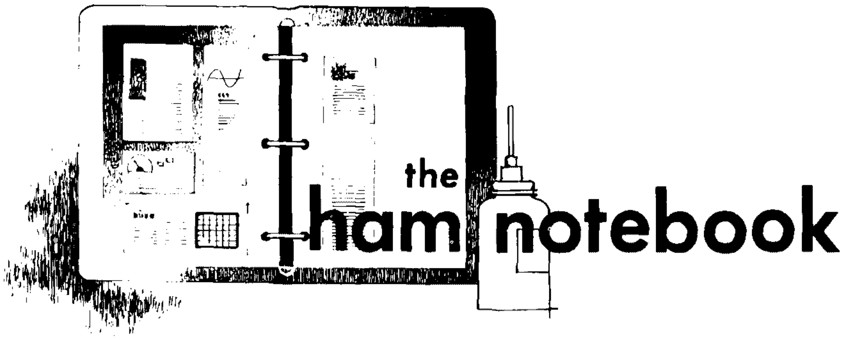



## simple filter reduces frequency synthesizer sidebands

In the past few years frequency synthesizers have become popular for vhf and hf use. Most follow the basic scheme of fig. 1 and use the Motorola 4044 (or Fairchild 11C44) phase detector IC. While good results are fairly easy to obtain, many designs miss a trick by simply copying the filter circuit in the data sheets. As a result, too much reference frequency feeds through and undesirable fm sidebands appear in the output.

As shown in fig. 1, the vco tunes from 7 to 8 MHz as its control voltage goes from 8 volts to 4 volts. The control

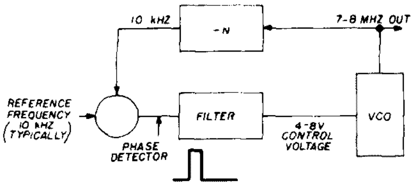



**fig. 1. Frequency synthesizer loop diagram showing relationship of basic components.**

voltage is only an amplified and filtered version of the phase detector output (ideally it is pure dc). A change in frequency over this range requires that the phase detector provide more or less dc accordingly. The trouble is that the phase detector outputs a *pulse train and* you are interested only in its *average* dc value (it is rather like a switching regulated power supply). If you filter very heavily, you will get pure dc but the control voltage will respond too slowly to be useful. Notice that if the vco is right on frequency, the phase detector

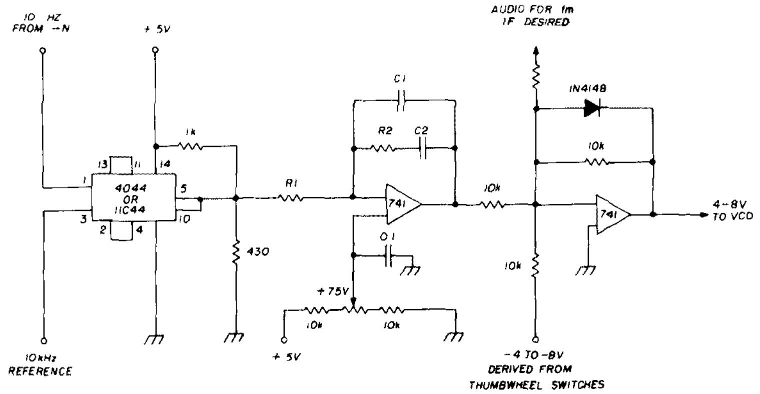



**fig. 2. Schematic diagram of improved frequency synthesizer with filter circuit using readily available components, and featuring greater than 60 dB suppression of sideband frequencies.**

must only provide a very small correction voltage. Clearly, the answer is to make the phase detector do as little work as possible. Since it operates by putting out pulses of constant amplitude and varying duty cycle, you want to reduce the pulses to nearly zero width.

The first trick is to keep the phase detector from doing all the work itself. While digital thumbwheel switches are used to set the frequency divider, the same digital information can be used (via a simple D-to-A converter) to provide some of the vco control voltage that will be needed. The other trick is to find a way to get more dc gain without heavier ac filtering. Discard the cheap *and easy circuitry provided on the phase* detector chip in favor of the slightly more complex arrangement shown in fig. 2. Note that at dc, the first 741 op amp runs at full throttle (at gain of 100,000), with the same filtering as the former circuit. Components can be chosen by the formulas provided for good performance with reference frequencies greater than 5 kHz, and LC oscillators for the vco.

$$R1 = \frac{1}{4.7 \times 10^6} \cdot \frac{K_o}{nC1} \text{ (ohms)}$$

$$R2 = \frac{1}{1350 \times C1} \text{ (ohms)}$$

$$C2 = \frac{C1}{14} \text{ (farads)}$$

where $K_o$ is the vco characteristic (in Hz/V) and should be approximately 200 kHz per volt, *n* is the middle-of-the-range divider reading, and C1 (farads) is chosen to provide a value of R1 of approximately 10k ohms.

It is really worth the cut-and-try to get a reasonable linear frequency output vs control voltage characteristic, as this results in sideband reduction. A very abrupt varicap can help here. I have had good luck with the KV5001 ($.70 from Solid States Sales, Somerville, Massachusetts).

To derive a tuning voltage from the thumbwheel switches you can buy a

good $12 D-to-A converter from Hybrid Systems, Burlington, Massachusetts or make your own from cmos gates and some resistors (as shown in the RCA COSMOS circuits manual).

Tuneup is simple. Set the synthesizer to the middle of its range and observe the phase detector output on a good fast oscilloscope. You should see a dc level of about 1.5 volt with a string of needle-narrow pulses shooting either up to 2.25 volts or down to 0.75 volt, adjust the pot until they nearly vanish or are up half the time and down the other. Sidebands should now be more than 60 dB below the carrier.

**Eric Blomberg, K1PCT**

## improved reliability for Collins KWM-2 transceivers

During the past fifteen years Collins has sold over 27000 KWM-2 transceivers, an impressive testimonial to the reliability and conservative design of this piece of equipment. However, minor changes to the KWM-2 over the years have been made to bring the design to its present state; and one of these changes is important to owners of older-model KWM-2s.

Connoisseurs of Collins equipment are quick to point out that KWM-2s with the new "meat ball" emblem on the panel are better than the older models which sport the "wing" emblem. The "wing" models are alleged to have "relay" problems, solved in the later models by incorporation of encapsulated plug-in relays.

In my opinion this subtle distinction is not valid, inasmuch as the alleged relay problems don't seem to have anything to do with the *type* of relays installed, and contact burn (normally avoided by encapsulated relays) is not a problem. Instead, the difficulty appears to be that the vox relay (K2) coils in earlier model KWM-2s sometimes have a tendency to burn out, whereas later models do not have this problem. And a vexing problem it is, too, as anyone who has attempted to rewire the vox relay in a KWM-2 can tell you! It is an onerous, time-consuming, exacting, and humbling job.

A much easier approach is to modify the KWM-2 relay coil circuit to prevent coil burn-out, a modification recommended to owners of older KWM-2s that may not have already incorporated this simple change.

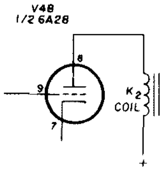



**fig. 3. Original vox relay circuit used in the Collins KWM-2 transceiver, without protective resistor.**

The vox relay circuit for older KWM-2 models is shown in **fig. 3**. The vox relay actuator tube (V4B) operates under vox control, push-to-talk, and when the transceiver is actuated for CW. When the vox circuit is actuated, the relay tube conducts heavily and plate current flows through the coil of vox relay K2. In this design, the plate current of V4B is an order of magnitude greater than the maximum specified relay coil current. Overheated relay coils have a tendency to burn out, and the relay in the KWM-2 is no exception.

In some models of the KWM-2 a protective bias resistor of 330 ohms is connected between the cathode of the vox relay amplifier tube V4B (pin 7) and the tip contact of the microphone jack. The presence of this resistor can be determined with an ohmmeter. If present, the following description for adding a 12k, 2-watt resistor in the plate circuit may be disregarded.

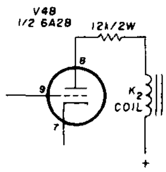



**fig. 4. Simple modification of the relay circuit to provide protective 12k resistor between plate of tube V4B (pin 8) and coil of the vox relay.**

The circuit modification is simple (**fig. 4**). A 12k, 2-watt resistor is added in series with the relay coil to reduce the coil current to a safe value. This addition in no way affects relay operation.

You can check your KWM-2 to find out if this modification has already been included in the circuit. In most models, the resistor is connected from pin 8 of tube socket XV4 (6AZ8) to an adjacent terminal point. If this resistor or the cathode resistor is missing, the following modifications are suggested:

1. Disconnect the red-white wire from pin 8 of socket XV4 and connect it to a terminal tie point. It is possible to mount a small standoff terminal on a bolt of the socket, or else to epoxy a miniature terminal strip adjacent to the socket.

2. Install a 12k, 2-watt resistor (R202 in the latest KWM-2 circuitry) from the terminal point to pin 8 of socket XV4. This completes the modification. Apply power and test the vox circuit for activity.

**William I. Orr, W6SAI**

## Heath intrusion alarm for mobile equipment

The Heathkit GD-39 ultrasonic intrusion alarm should be ideal for mobile operators who want to protect their equipment. While I haven't made a battery-operated vehicle installation, the facts speak for themselves: 12V operation, 17.5V out of the rectifiers into the regulator, 13 mA current drain while waiting, and room on the back panel to mount some connectors (**fig. 5**). Some modifications will have to be made: a relay is required for spst operation (instead of switching the line), and two holes must be drilled into the board to bring in the battery supply.

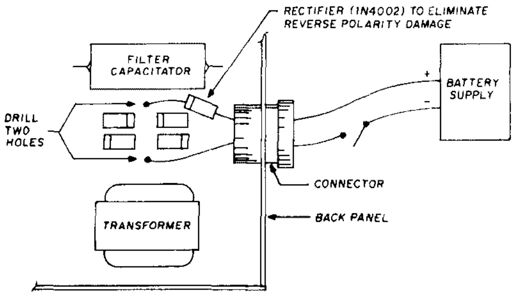
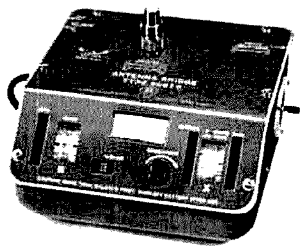



**fig. 5. Adaptation of Heathkit GD-39 intrusion alarm for mobile equipment.**

**William B. Rossman**

# NEW products

## Millen 90673 antenna bridge



The Millen 90673 Antenna Bridge is a radio frequency impedance measuring device which is intended for use in the frequency range from 1.8 to 30 MHz. An "extra" provided by this bridge is the measurement of rf resistance outside the normal high-frequency limit. Using additional coils, specially made for the vhf range, it is possible to obtain approximate resistance readings up to 60 MHz, but without any accompanying reactance calibration.

A null detector, with diode rectifier and dc amplifier, is built in and indicates on a 1-mA meter. The bridge requires an external oscillator, such as a grid-dip oscillator.

Resistance measurements are made by varying a differential air dielectric variable capacitor. The resistance range covered is approximately 5 to 150 ohms. At frequencies below 15 MHz, values up to 200 ohms are measurable. The accuracy of resistance measurement is better than ± 10 per cent over most of the range.

Reactance is measured by tuning a coil-and-capacitor circuit, first to an initial null with a resistive reference and then to a new null with the unknown impedance. This requires simultaneous balancing of resistance and reactance controls. The reactance range is calibrated in ohms of reactance at 1MHz and indicated inductive and capacitive values. The reactance value at the operating frequency is obtained by dividing the indicated reading by the frequency in use.

The Millen 90673 Antenna Bridge is based on the "Macromatcher" described in *QST*, (January, 1972). The circuit has been modified and the mechanical layout arranged to produce the commercial product. Since the original Millen 90672 Antenna Bridge was the starting point for the Macromatcher, the new 90673 Antenna Bridge is a logical and useful further development. The bridge is approximately 8 x 8 x 5 inches (20x20x13cm), not including plug-in coil storage box.

The 90673 Antenna Bridge normally measures coaxial-line connected impedances (unbalanced). A set of balun coils, with series and parallel reactances tuned out for amateur bands, is available at extra cost to measure balanced-line systems.

## power amplifier and preamp for 2-meter hand-helds

Hamtronics recently announced a new PA/Preamp Unit, the T130, for use with two-meter hand-held units and other low-power transceivers in either mobile or base-station applications. The *T130 PA/Preamp Unit* provides up to 25-watts output for inputs between 200 milliwatts and 6 watts (specify output power of your transceiver).

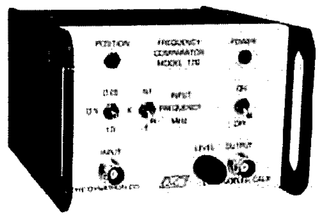
The power amplifier section is solid state, requires no tuning, and the preamplifier portion provides 20-dB gain, using a low-noise fet cascode preamp circuit. Transmit-receive switching is automatic, and a front-panel switch allows the unit to be turned off, and the transceiver bypassed to the antenna, when fringe-area coverage is not required.

The PA/Preamp unit is attractively packaged in a high-impact *Cycolac* case with brushed-aluminum panels. Your hand-held connects to a front-panel BNC connector, while the antenna is connected through a rear-panel uhf connector. *On/off* status is indicated by an LED. In operation, the current drain is 4 amperes.

If desired, the unit is also available with a front-panel loudspeaker and audio amplifier to provide improved receiver audio for your transceiver. Another option diverts the transceiver's audio to your car's a-m broadcast receiver, for the same purpose. The 3-1/2 by 7-1/2 by 7-inch (9x19x18mm) cabinet provides ample room for your own additions; *e.g.*, microphone connector and hanger, a clip to hold your hand-held unit, and a 12-volt lead for its operation.

The basic T130 PA/Preamp Unit is $129.95. A version without the preamplifier is available for $109.95. For a complete catalog of vhf and uhf products, send a self-addressed, stamped envelope to: Hamtronics, Inc., 182 Belmont Road, Rochester, New York 14612.

## frequency comparator



The Dynatron Company has announced a new frequency comparator, the *DyCo Model 176*, intended for use in calibrating crystal oscillators and time bases in digital frequency counters, communications monitors, and frequency meters against television network atomic frequency standards.

The calibration scheme, devised by The National Bureau of Standards, can now achieve in minutes the same results that previously took hours by receiving WWVB. Traceability to NBS is assured by the monthly publication of the network offset frequencies in the *NBS Services Bulletin*.

A color TV receiver (any set will do) receives the network color subcarrier and displays the result of the comparison as a vertical rainbow bar. Frequency calibrations to accuracies within parts in $10^{10}$ take just a few minutes.

The *DyCo Model 176* accepts inputs of 0.25, 0.5, 1.0, 2.5, 5.0, or 10 MHz. Price of the *Model 176* is $179.95. For additional information, write The Dynatron Company, Box 48822, Los Angeles, California 90048.

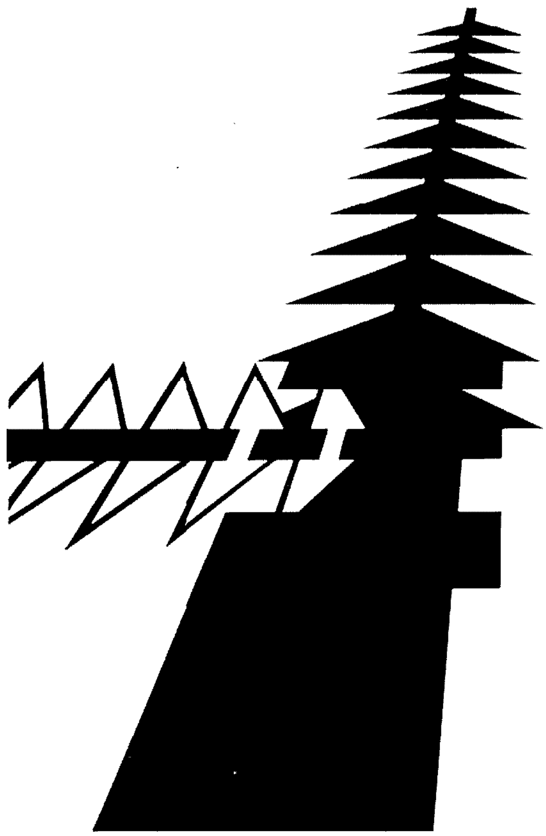
one dollar

# ham radio magazine

JULY 1977



## 1296-MHz TRANSVERTER



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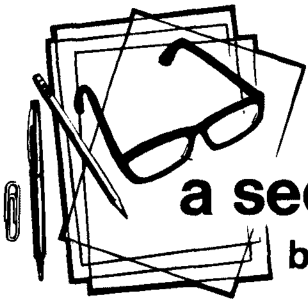


# a second look

**by Jim Fisk**

The editor of a technically oriented magazine like *ham radio* has to wear several hats. I could use this whole page to describe the details that require attention to keep the magazine running smoothly, but I'd like to talk for a moment about one very important editorial task that means the difference between an interesting, accurate technical magazine, and one that isn't.

The articles in *ham radio are* written by authors not unlike you, the reader. They range from enthusiastic hams who want to share an idea, to amateurs with engineering backgrounds (who also want to share an idea). I welcome the output of anyone who is interested in contributing something which will benefit all hams.

Budding authors often ask, "What kind of articles are you looking for?" That question is difficult to answer since many new manuscripts arrive every day, but generally speaking, I am looking for simple construction projects that the average reader can put together in one or two weekends. Larger construction projects are also welcome, but the average *ham radio* reader must split his leisure time between amateur radio and other interests, so he doesn't have time to build a Chinese copy of a complex piece of gear.

When I read an article contributed to *ham radio,* the first thing I look for is interest value. If the manuscript passes this test, the next thing I look for is technical accuracy and attention to detail. The contributed article doesn't have to be a literary masterpiece. If you have a good idea; if it's well documented; if the illustrations and technical discussion are clear and accurate — you may have a winner!

Don't feel too badly if your article is not accepted. Since we publish about twelve feature articles in every issue, to keep the production pipeline full we purchase that same number of new manuscripts each month. This keeps the article backlog to a minimum and insures that the fare served up in each issue of *ham radio* represents the latest possible information on any given subject.

During an average month I receive 50 to 60 new manuscripts; at some point during the month I sit down and go through each of the articles rejecting those that are clearly not usable in *ham radio*; this narrows the stack of manuscripts down to 25 or so.

Now comes the most difficult part — deciding which of those remaining 25 articles are most desirable. Articles that are too long or need more polishing are returned to the author with suggestions for rewriting the article to our standards. The remaining material is further screened for technical accuracy and reader interest; this process continues, with comments from the staff (and usually arguments), until we have decided upon the articles which will be published in *ham radio.*

If you prefer to read *ham radio,* rather than write for it, you can help by telling me the kind of article you like. If you have a pet project in mind, or an old project that could be updated with transistors or ICs, let me know about it; I'll pass the idea along to one of our regular contributors.

## *1977 ham radio sweepstakes*

The winner of the Grand Prize in the 1977 *ham radio sweepstakes,* a Kenwood TS-820 six-band transceiver, one of the most exciting of this year's new amateur transceivers, *and* the versatile Kenwood TR-7400A two-meter fm transceiver, is Clarence "Bick" Bickford, KØRHP. Other happy winners in this year's Sweepstakes are Bill Locke, W4RPU, and George Gruetzmacher, WB9QBA, who each won TS-820 transceivers; and Hal Tune, W8LZE, and Jack Langley, KØMER, who are proud new owners of TR-7400A fm rigs.

This year's sweepstakes was one of our biggest ever, and we want to thank all of you who took the time and effort to enter. Next year we'll be doing it again, so stay tuned . . .

**Jim Fisk, W1HR**
**editor-in-chief**

# presstop

A NEW "13-METER" BAND, 15 meters back where it belongs, and 40 meters exclusively Amateur were the big pluses in the FCC's WARC Fifth Notice of Inquiry, released May 20th. On the negative side, the proposed 160-190 kHz band has been dropped entirely due to objections from the power-generating industry which uses those frequencies for carrier-current telemetry and control of their high-voltage transmission lines; they fear strong Amateur signals could cause interference problems.

Looking At The Latest proposals band-by-band and referencing them to last December's Third Notice of Inquiry (February HR) finds:

| | |
|---|---|
| 1600 Meters: | 160-190 kHz dropped. |
| 160 Meters: | 1750-1800 kHz dropped; 1800-1900 kHz exclusive; 1900-2000 kHz restored, shared with existing services (Loran). |
| 80 Meters: | 3500-3900 kHz exclusive; 3900-4000 kHz shared (no change). |
| 40 Meters: | 6950-7250 kHz exclusive; 7250-7300 kHz dropped. |
| 20 Meters: | 13950-14400 kHz exclusive (no change). |
| 15 Meters: | 20700-20950 kHz dropped; 20950-21200 kHz exclusive; 21200-21450 kHz exclusive (restored). |
| 13 Meters: | 25.76-25.86 MHz new, exclusive. |
| 10,6,2 and 1¼ Meters: | all remain as they are now and were proposed in December. |
| 3/4 Meters: | 420-450 MHz unchanged, except Amateur Satellite has been proposed worldwide for the 435-438-MHz slot. |
| 35cm: | 902-928 MHz shared unchanged. |

In The 1215-MHz Band, Amateurs would lose 1215-1240 MHz with 1240-1290 MHz shared with radiolocation and Amateur satellite sharing 1290-1300 MHz. The only other change in the higher frequencies is a partial restoration of the previously dropped 48-50 GHz band to 49.8 to 50 GHz.

It Must Be Emphasized that these bands are only proposed and only for the U.S. position at the 1979 World Conference. They'll be reviewed by the Advisory Committee on Amateur Radio, tentatively scheduled to meet in Washington on July 12. Loss of the 160-190 kHz band will certainly be discussed, and perhaps another spot in that spectrum can be found for us. The new 25.76-MHz band, a useful addition, is not nearly so significant to Amateur needs as the 10- and 18-MHz bands we previously requested, so the ACAR should be focusing on that area, too.

Comments On This Fifth Notice of Inquiry are due August 1st, and Reply Comments August 22nd. We'll need lots of help if we're to preserve the relatively good position we're now in.

GETTYSBURG'S SPECIAL LICENSING CHIEF was indicted April 29th by a Harrisburg, Pennsylvania Federal Grand Jury on four counts of bribery. The indictment charged FCC official Richard C. Ziegler with the alleged solicitation of $100 each from four individuals in exchange for influencing the processing of their callsign applications last spring, according to the report in the April 29 Harrisburg Patriot. The indictment apparently was the result of an FBI investigation begun last fall.

220-MHZ CLASS-E CB still seems to be a live issue with the FCC, according to a report from the Commission's Office of Plans and Policy that was released in late April. The document, "Spectrum Alternatives of Personal Radio Service," was prepared by the Personal Radio Planning Group, which considered 17 alternative spots between 25 and 1215 MHz for personal radio expansion.

Based On Such Factors as user loading, costs of relocation, and TVI potential, seven segments were selected: 26.95-26.96, 27.54-28.00, 29.80-29.89, 29.91-30.00, 222-224, 894-902, and 928-947 MHz. Since many of the problems of the present Class-D band seem to rule out any of the slots in the 26-30-MHz region, we're back to that same old basic conflict between 220 and 900 MHz. However, despite the Planning Group's determination that its selections were the only likely current candidates, their report did not preclude consideration of some totally different spot, and even suggested that the Commission could decide that further CB expansion was not in the public interest.

EAST COAST VHF/UHF CONFERENCE in New Hampshire May 7th and 8th was rated a "huge success" by the nearly 100 attending from the U.S. and Canada. Noise figure competition winners were K2UYH with his V-244 based 432-MHz preamp at 0.95 dB, and W1JR's 144-MHz design with 0.8 dB.

Ham Radio's Annual Award for Technical Achievement went to the Mt. Airy VHF Club for the Pack Rats' 432-MHz EME DXpedition to Colombia last summer — W3HMU accepted the plaque for the club.

ARMY RESERVIST AMATEURS can get their two-week summer duty teaching teenagers Amateur Radio in the FAA's Career Interest Program. For details call Col. Colby or Lt. D'Angelo at (202)325-8483, or write 118 South Royal, Alexandria, Virginia 22314.

# 1296-MHz transverter

Complete construction details for a simple, inexpensive transverter for ssb and CW that will make a noticeable dent on the 1296-MHz amateur band

**Recent issues** of *ham radio* and other amateur publications have contained a wealth of construction articles on equipment for 1296 MHz, indicating the growth of interest and activity on that band. Conspicuously absent from the literature, however, is a simple, inexpensive way of generating reasonable amounts of stable transmitter output power — greater than 1 watt — for serious CW and ssb work on 1296 MHz.

Most long-haul, narrow-band DX work is conducted at or just above 1296 MHz, leaving the lower part of the band for wideband modes. Traditional transmitting schemes for this band usually involve tripling from 432 MHz using planar triodes in a cavity or stripline arrangements, and more recently, varactor diodes. These approaches yield CW or fm signals, but they are obviously unsuitable for single sideband.

Recent solid-state mixer designs, although suitable for producing clean ssb and CW signals have one principal drawback: low power output, typically in the dozens of milliwatts range. This requires considerable linear amplification to approach reasonable power levels. The circuitry associated with uhf mixers is difficult for amateurs unfamiliar with these devices and expensive in terms of dollars invested per watt of outpuut power obtained.

In an effort to overcome these drawbacks with minimum circuit complexity and financial investment, a high-level mixer was developed which uses the popular 2C39/7289/3CX100A5 family of planar triodes. These tubes are abundantly available surplus at extremely reasonable prices. Depending on the plate voltage applied to the tube, this transverter will deliver from 5 to 15 watts of clean, stable CW and ssb power output on 1296 MHz. This is more than enough for routine contacts up to a 100 miles (160km) or more. My 1296-MHz signals are regularly copied at +20 dB over S-9 over a 50 mile (80km) path using the transverter stage alone; for more serious DX work the unit will drive a single 7289 to 100 watts ssb and CW output!

## theory of operation

The 1296-MHz transverter operates like a receiving converter or mixer in reverse, and at much higher power levels. As shown in **fig. 1**, an ssb signal from the output of a high-frequency or vhf transmitter (here considered to be the *intermediate frequency* or i-f) is mixed with a higher frequency carrier (the local oscillator or LO) to produce sum and difference frequencies, of which one is the desired uhf ssb signal. The remaining, undesired signals are eliminated with a selective filter.

I used the output of a 50-MHz ssb transceiver for my i-f. Obviously, other ssb source frequencies could be used, but it is desirable to use as high an i-f as possible to separate the desired mixer product from the unwanted LO and difference frequencies as much as possible, making it easier to eliminate the unwanted signals by filtering. Intermediate frequencies as low as 21 or 28 MHz can be used with little difficulty.

Transceivers in the 10-20 watt class are ideal for driving this transverter. They should, however, be

**By Joe M. Cadwallader, K6ZMW,** Star Route 2, Bosse Road, Jackson, California 95642

isolated from the transverter by a simple 3 dB attenuator[1] (such as a suitable length of RG-58/U coaxial cable) to make sure the transceiver is terminated in a matched, resistive load. The output of transceivers in the 100-watt class should be attenuated down to about 10 watts; don't just turn down the DRIVE or MIC GAIN control.

The transverter requires about 5 watts of local oscillator injection at 1296 MHz plus or minus the i-f. This signal can be derived in a number of ways. In my case, withh a 50-MHz i-f, a LO of either 1246 or 1346 MHz was needed. A crystal-controlled signal source providing about 10 watts output at 415.333 MHz was built; this was used to drive a tripler stage to 1246 MHz with about 5 watts output.

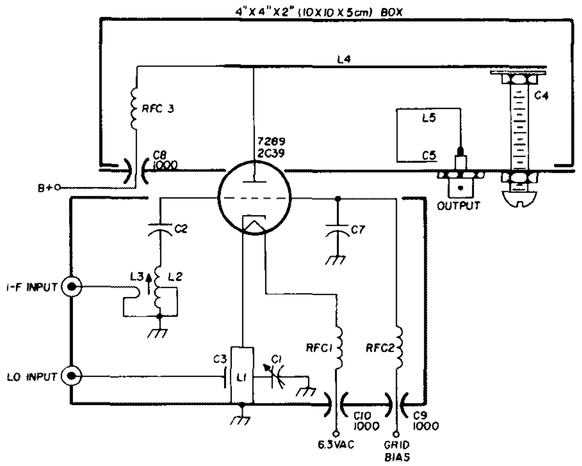
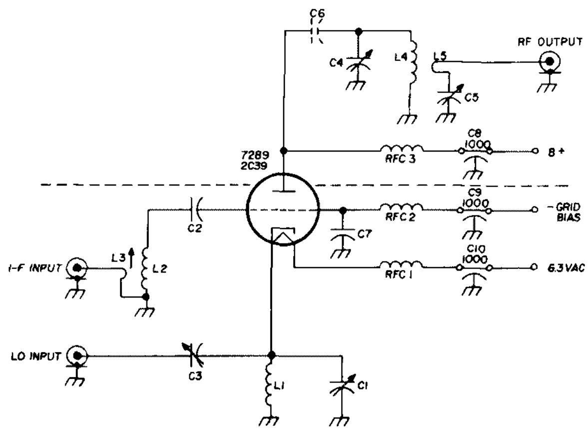



**fig. 1. Circuit layout for the 1296-MHz transverter. Component details are listed under fig. 2.**

There are a number of ways to generate the 415-MHz signal: the easiest is to modify and retune an existing transmitter which operates near this frequency. Many 432-MHz transmitters described in amateur publications can be easily retuned; transistorized transmitter kits advertised in amateur publications are very reasonably priced and should work well for this purpose.

Even old commercial 450-MHz fm transmitter strips, often available as junk, work nicely. If this approach is used, however, a few precautions are in order: turn the DEVIATION control off and, if possible, remove the speech-amplifier and phase-modulator tubes; substantially increase the power supply filtering to assure clean output with no ac hum which would otherwise appear on your transverter LO signal; voltage-regulate the oscillator and buffer stages with zener diodes or VR tubes to maintain oscillator stability; and use a good quality crystal with a low temperature coefficient in a temperature-controlled oven, or mount the crystal under the




| | |
|---|---|
| C1, C3 | part of cathode circuit (see fig. 5) |
| C2 | 150 pF silver-mica capacitor for 50-MHz i-f (three 47-pF dipped silver-mica capacitors in parallel) |
| C4 | plate tuning capacitor (see fig. 7) |
| C5 | part of L5 (see fig. 3) or 10 pF piston trimmers |
| C6 | non-existent; represents dc open condition of this line configuration (see text) |
| C7 | grid bypass capacitor (see fig. 4) |
| C8,C9,C10 | 1000 pF feedthrough capacitors. C8 must be rated for applied B+ voltage |
| L1 | part of cathode circuit (see fig. 5) |
| L2 | 4 turns no. 16 (1.3mm) enamelled copper wire on 1/4" (6.5mm) slug-tuned coil form (for 50-MHz i-f) |
| L3 | 1 turn no. 16 (1.3mm) around cold end of L2 |
| L4 | plate line (see fig. 6) |
| L5 | 1/4" (6.5mm) wide copper or brass strip, about 1/8" (3mm) away from plate line (see fig. 3) |
| RFC | 15 turns no. 16 (1.3mm) copper wire, closewound on 1/16" (1.5mm) mandrel |

**fig. 2. Schematic diagram of the 1296-MHz transverter. The circuit layout is shown in fig. 1. Rf power output at 1296 MHz is 17 watts.**

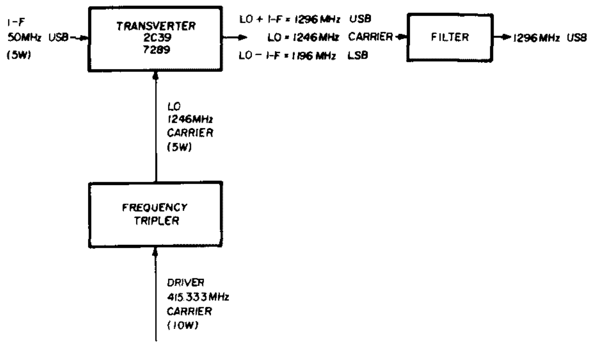
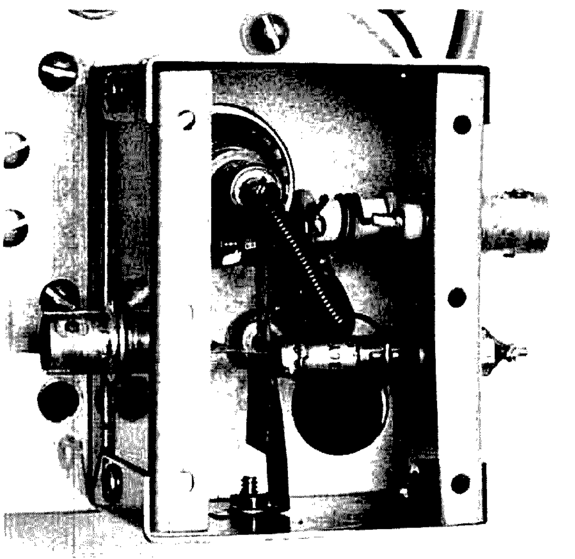



**fig. 3. Block diagram of the 1296-MHz transverter system for CW and ssb operation. Although the author used a 50-MHz ssb/CW transmitter, 21 or 28 MHz could be used with equally good results. Frequencies below 21 MHz are not recommended because of the difficulty in separating the resulting mixer products.**

chassis away from sources of heat, to reduce LO drift.

Regardless of which approach is used to generate the LO signal, a small amount can also be coupled off for LO injection to the receiving converter, thus reducing the total system equipment requirement and yielding true transceive operation on 1296 MHz. In my case, a small amount of 415.333-MHz energy was inductively coupled from the PA grid circuit of a 10-watt transmitter strip used as the LO source and applied to the multiplier diode of a popular trough-line receiving converter.[2]

The task of tripling up to the transverter's required LO injection frequency can be readily accomplished in a varactor multiplier — either commercial* or homebrew[3] — or a stripline or cavity multiplier stage[4,5] can be built around a 2C39/7289 triode. It has been suggested that cavity assemblies from surplus uhf equipment such as the UPX-6 would serve this purpose well. As is, these beautiful cavities tune from roughly 1000 to 1200 MHz.

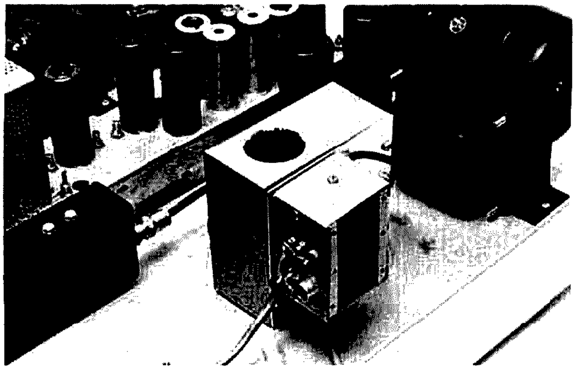


Close-up of the cathode compartment. The BNC connector on the left is used for the local oscillator input, while the one on the right is for the i-f input. The filament and grid voltages are brought into the compartment by feed-through capacitors.

Table 1 lists the required local-oscillator frequencies for various intermediate frequencies. Keep in mind that using an LO *above* 1296 MHz causes inversion of the sideband in the transverter: for example, a 1346-MHz LO minus a 50-MHz *upper* sideband signal equals 1296-MHz *lower* sideband.

Referring to the schematic diagram, **fig. 2**, the

*The MMv-1296 tripler available from Spectrum International, Box 1084, Concord, Massachusetts 01742.



The complete 1296-MHz ssb/CW transverter, mounted on a pressurized chassis. The local-oscillator chain is at the upper left, the varactor tripler is to the left, and the high-voltage power supply and blower are at the right.

transverter circuit is remarkably simple, using a 7289 (2C39) or equivalent in the familiar grounded-grid configuration. As is common practice in this application, the tube grid is not actually grounded directly but rather is bypassed, through capacitor C7 so the grid is grounded at the signal frequency while remaining above ground to dc. This provides a convenient way to apply grid bias through RFC2 and C9 without affecting the rf behavior of the grid circuit.

**table 1.** Possible i-f/local oscillator combinations for the 1296-MHz transverter. LO frequencies above 1296 MHz invert the sideband.

| intermediate frequency | local oscillator | driver (LO ÷ 3) |
|---|---|---|
| 21 MHz | 1275 MHz | 425.000 MHz |
| 21 MHz | 1317 MHz | 439.000 MHz |
| 28 MHz | 1268 MHz | 422.666 MHz |
| 28 MHz | 1324 MHz | 441.333 MHz |
| 50 MHz | 1246 MHz | 415.333 MHz |
| 50 MHz | 1346 MHz | 448.666 MHz |
| 144 MHz | 1152 MHz | 384.000 MHz |
| 144 MHz | 1440 MHz | 480.000 MHz |

The grid bypass capacitor, C7, consists of a flat, concentric brass or copper plate connected by finger stock to the tube's grid collar and insulated from the chassis with a thin mica or Teflon sheet. This type of bypass plate is standard equipment on military surplus vhf communications gear and can often be scavenged. The bypass plate can also be home built by soldering finger-stock material* around an appropriately sized hole centered on a flat brass or copper sheet (**fig. 4**). Thin mica for the dielectric material is available in most hardware or plumbing supply stores as replacement material for gas furnace pilot-light inspection holes. It can be easily cut with scissors or an *Exacto* knife.

The bypass plate is insulated from its mounting

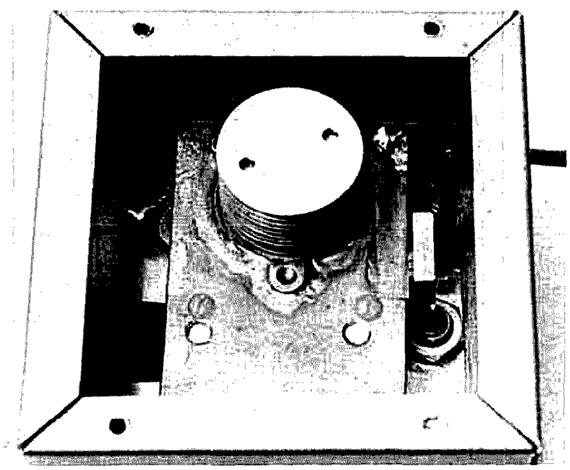
*Instrument Specialties Company, Little Falls, New Jersey 07424.

screws with the same type of nylon bushings which are used to mount and insulate power transistors. Once assembled, a typical value for this bypass capacitor is about 100 pF; this represents about 1 ohm of capacitive reactance — essentially a dead short — at 1296 MHz, and effectively grounds the grid at that frequency. However, at lower frequencies the grid is definitely *not* at rf ground: at 50 MHz the reactance of the grid bypass is about 30 ohms, and at 28 MHz it is about 55 ohms. Thus, the grid can be driven by the low-frequency ssb i-f signal while the cathode is driven by the high frequency LO signal: the sum and difference of these two signals appear in the plate circuit.

This simple approach can be applied to numerous uhf tubes in various grounded-grid configurations, stripline or cavity, commercial or military surplus, with equally good results.

The cathode circuit, **fig.** 5, driven at the LO frequency of 1246 MHz in my case, consists of a shorted quarter-wave line section L1, made of thin brass or copper sheet 1/4 inch (6.5mm) wide, wrapped around the tube cathode sleeve and running 1-3/4 inch (44mm) to chassis ground. The line is tuned with C1 which may be a low loss, high quality glass or ceramic piston trimmer or, better yet, a metal tab bent up near the line, or a brass machine screw with a brass disc about 1/2 inch (13mm) in diameter soldered to its end (similar to C4). The LO energy is capacitively coupled to the middle of this line by C3, a small brass or copper tab soldered to the LO input connector and bent up near the cathode line. Spacing can be adjusted for maximum LO drive.

The ssb i-f signal is coupled to the grid circuit by L3, a one-turn link wound around the cold end of L2. The L2-C2 circuit must resonate at the intermediate frequency. Since this circuit will be driven with about



The plate-line enclosure for the 1296-MHz transverter. The output coupling network, L5-C5, is to the right.

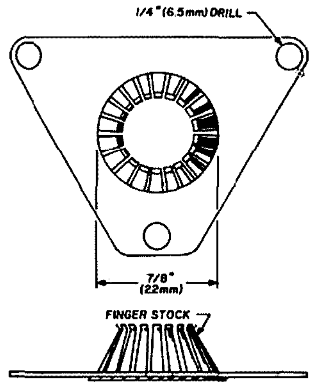
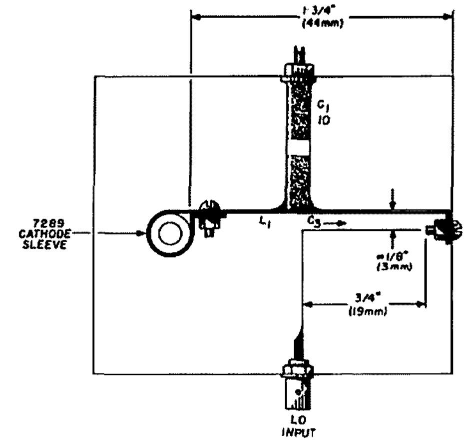



**fig. 4. Construction of the grid bypass capacitor, C7.**

5 watts, these coils should be either air-core or wound on a low-loss slug-tuned ceramic coil form of moderate diameter (1/4 inch [6.5mm] minimum, larger preferred) using no. 14 (1.6mm) to no. 18 (1mm) enameled copper wire. C2 should be a good quality mica or silver-mica capacitor to minimize losses. Two or three capacitors may be paralleled to handle the required rf current and still resonate with L2 at the i-f. The C2-L2 combination should be pre-

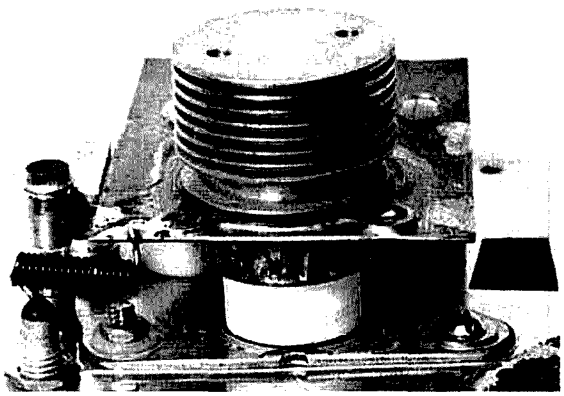



| | |
|---|---|
| C2 | 10 pF piston trimmer capacitor or built as C4 (fig. 7) |
| C3 | 1/4" (6.5mm) wide copper or brass strap |
| L1 | 1/4" (6.5mm) wide copper or brass strap |

**fig. 6. Construction of the cathode circuit for the 1296-MHz transverter. This circuit is installed in a small aluminum minibox which is mounted on the plate enclosure (see fig. 1 and the photographs).**

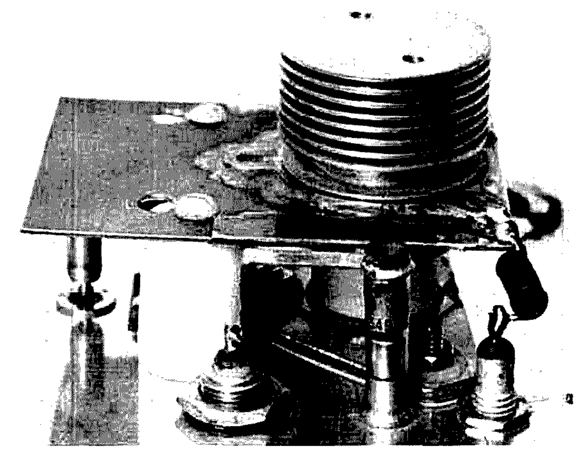
tuned to the intermediate frequency with a grid-dipper before installation.

The plate circuit (fig. 1) consists of an open half-wave line section, L4, tuned to 1296 MHz by capacitor C4. The line section is made of sheet copper or brass with finger stock connections to the tube anode ring and supported by insulating columns of Teflon, Rexolite, ceramic, or other good uhf dielec-



Construction details of the plate line, output network, and mounting of the grid-bypass capacitor, capacitor C7. Note that the mounting screw on the left is connected to the grid bypass plate for grid bias (from the cathode enclosure, below).

tric. The anode finger stock assembly can be found in the same military surplus vhf communication gear as the grid bypass assembly. Alternatively, commercial finger stock may be formed to the appropriate size and soldered around a hole in one end of the



Another view of the plate line showing the grid bypass capacitor, C7; output link, L5-C5; plate rf choke, RFC3; and bypass capacitor, C8.

plate line, L4, large enough to accommodate the tube anode ring as shown in **fig. 6**.

Capacitor C6 does not really exist, but merely represents the dc-open condition of this line configuration. In fact, the entire line and tube anode have B+ applied through RFC3 and C8. This feedthrough bypass capacitor must be rated to withstand the B+ voltage applied to the tube. If a commercial or surplus bulkhead capacitor cannot be found, this capacitor can be home-made using a 1-1/2-inch (38mm) square of sheet metal and sheet mica.

The plate tuning capacitor, C4, requires about 10 pF and must not break down at full applied B+. A low loss, high quality glass or ceramic piston trimmer (rated accordingly) would do nicely, or this capacitor may be constructed using 3/4-inch (19mm) wide, sheet brass strap bent up near the plate line for about 3/4 inch (19mm) and insulated from it with a layer or two of sheet mica (see **fig. 7**). The spacing between this tab and the plate line can be varied by any convenient mechanical means to tune the line. An alternate approach is to drill and tap the transverter topplate to pass a no. 8 or no. 10 (M4-M5) brass machine screw with captive lock nut. Cut the screw so it is just short of touching the plate line (by at least 1/16 inch or 1.5mm) and solder a 3/4 inch (19mm) diameter brass disc to its end. Insulate the disc from the plate line with mica sheet.

Output power is inductively coupled from the plate line via L5 which is made from 1/4-inch (6.5mm) wide copper or brass strap soldered to the output connector (N or BNC type) and run parallel to the plate line about 1/8 inch (3mm) away. The end of this strap can then be run down to, and parallel with, the chassis to form the matching capacitor C5. A low-loss glass or ceramic piston trimmer of about 10 pF capable of handling some moderately large rf currents, could also be used.

## construction

The grid i-f and cathode LO circuitry plus filament and grid bias wiring are all contained within (and shielded by) a small aluminum minibox mounted on top of the plate line enclosure. This minibox is secured by the same hardware which is used to mount the grid bypass plate, C7. Two screws are insulated with nylon bushings from this grid bypass plate; the third screw makes contact with C7 but is still insulated from the chassis by nylon bushings. Bias and i-f connections are then made to this screw.

The plate circuitry and output coupling loop are

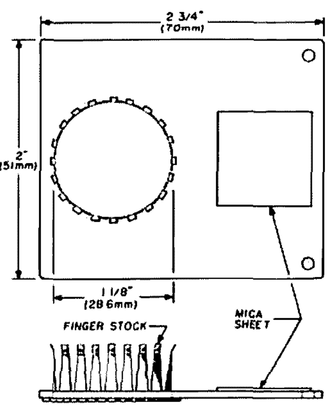
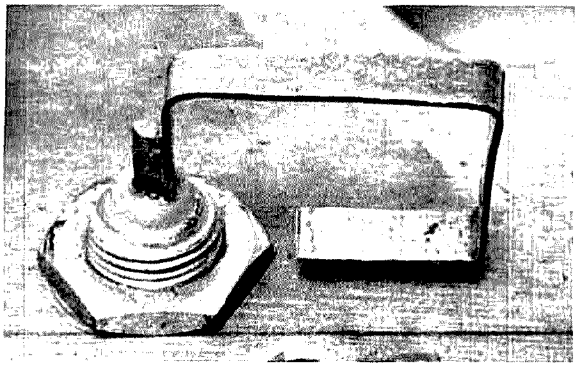



fig. 6. Plate line (L4 in figs. 1 and 2) for the 1296-MHz transverter. Finger stock can be obtained from Instrument Specialties, Little Falls, New Jersey.



An alternate arrangement for the output circuit. L5-C5.

contained within a 4 x 4 x 2 inch (10x10x5cm) minibox which serves as the chassis base. Screened air holes provide a path for forced air cooling of the tube anode structure, which is absolutely essential at reasonable power levels. These holes must be rf tight, so the shielding screen must be well grounded around its periphery.

Copper screen can be soldered to the aluminum

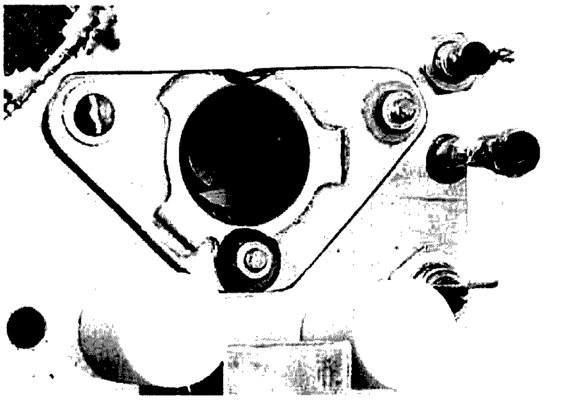


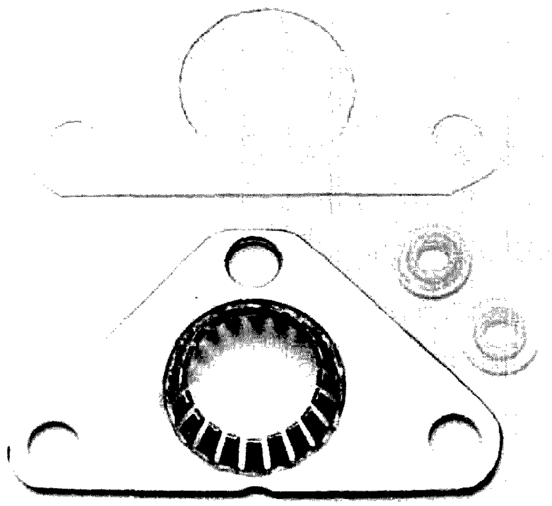
Plate line removed to show grid bypass plate. The plate line is mounted on the two Teflon pillars at bottom.

chassis by first tinning the aluminum with a large soldering gun or iron. Generously coat the periphery of the air hole with a heavy oil (clean auto engine oil works fine) to keep the aluminum from being exposed to air, then sand the surface clean using ordinary sandpaper or emery paper. Once cleaned, a hot iron (200 watts minimum) and rosin-core solder will tin the area beautifully. Then the copper screen can be soldered to the aluminum chassis. Practice this procedure first on some small aluminum scrap!

A convenient way to provide cooling air is to mount the transverter on edge on a large, air-tight pressurized chassis. The air hole in the transverter plate-line enclosure should be placed over an equal-sized hole in the pressurized chassis.

This chassis can also serve as a mounting platform and air source for the power supplies and LO chain components.

Common vhf construction practice should be followed throughout including short component leads, quality component selection, and rigid mechanical assembly. In one case, a 1/16-inch (1.5mm) thick, 4-inch (10cm) square brass plate was



Construction details of the grid bypass plate showing the sheet mica insulator cut out with an Exacto knife, and the insulating washers.

used as the top plate of the plate line enclosure with good results. It is a good idea to sand bare the adjoining surfaces of the plate line enclosure, top plate, and bottom plate, and use additional screws to assemble the top and bottom to assure good

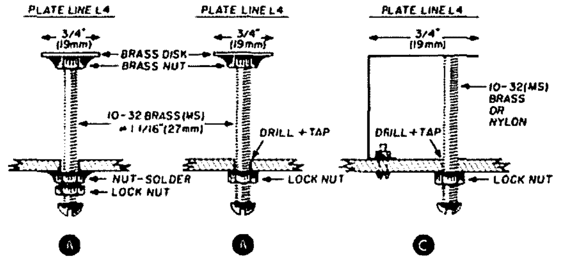



fig. 7. Three methods for building the plate tuning capacitor, C4.

shielding. When assembly is complete, a close visual inspection and a VOM continuity check should reveal any obvious problems before applying power.

## tuneup and adjustment

Initial testing can best be done using variable-

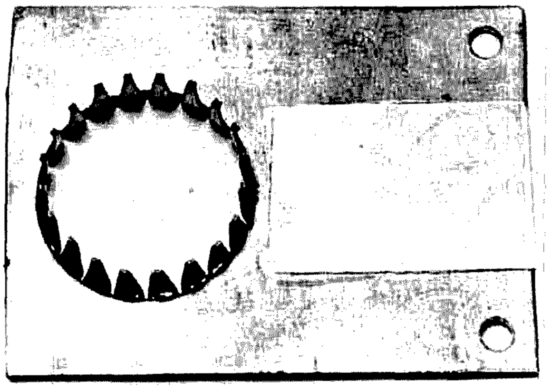


Plate line showing the plate collet (finger stock) and sheet mica insulator, right.

voltage power supplies to provide the plate voltage and grid bias. Since the grid will typically draw more than 50 mA, *the bias supply must have a low impedance and be capable of maintaining constant output voltage under fluctuating load conditions.* Apply filament voltage and cooling air, and after adequate filament warmup, gradually apply B+ while watching the plate current. Plate current should rise gradually, indicating normal tube conduction. Increase grid bias to reduce plate current until a plate voltage of about 300 volts can be applied with sufficient grid bias to limit plate current to about 20 mA.

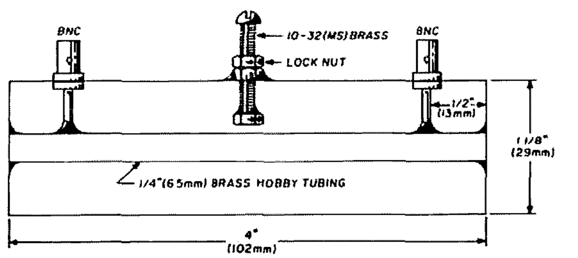



**fig. 8. Half-wavelength transmission-line filter for use with the 1296-MHz transverter.**

Under these conditions, gradually apply the LO drive to the cathode circuit. The presence of this signal will immediately be indicated by increased plate current. While not allowing plate current to rise above 100 mA, tune C1, C3, and the LO chain for maximum plate current. Increase grid bias as required to limit plate current to a safe value.

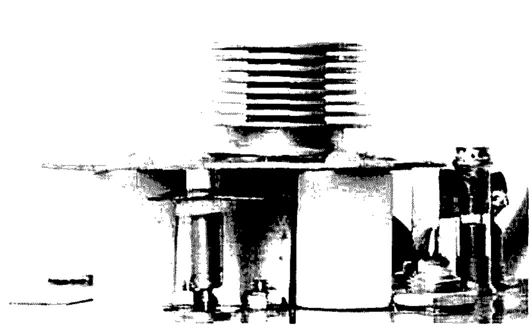
At this point, depending on your LO frequency, you *might* be able to resonate the plate circuit to the LO frequency by adjusting C4. Thus the unit operates as an amplifier to verify the behavior of the plate and output circuitry. If your LO is above 1296 MHz, tune toward minimum capacitance; if your LO is below 1296, tune toward maximum capacitance. Adjust C4, and the position of L5 and C5 for maximum indicated output power at the LO frequency.

Once cathode-line tuneup at the LO frequency has been accomplished and optimized, increase grid bias to reduce plate current to near cutoff — about 10 to 20 mA. Leave this grid bias at this value; the objective of this procedure is to bias the tube at or near cutoff (class AB or B) with plate voltage and LO drive applied but no i-f signal present. Now gradually apply a carrier at the i-f input and tune L2 for maximum plate current. Again, the presence of the i-f signal should immediately be indicated by increased plate current.

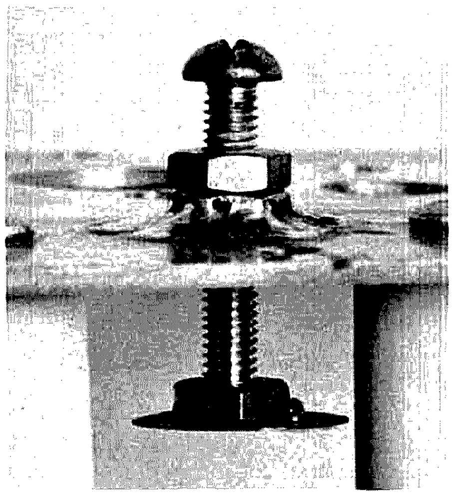
If all's well at this point, the LO and i-f signals are present in the tube's plate line; all that remains is to tune the plate line to the desired mixer product with C4. This process can be greatly simplified by placing a low loss filter tuned to 1296 MHz at the output of the transverter. An example filter of simple construction is shown in fig. **8**.[6]

The filter can be pre-tuned to 1296 MHz by placing it in series with your 1296-MHz receiving system and tuning for maximum received signal from a nearby 1296-MHz transmitter, the third-harmonic of a 432-MHz transmitter, the 9th harmonic from a 144-MHz transmitter, or any other convenient rf signal source. This assures that any transverter output observed with the filter present will be on the desired signal frequency, and that power measurements will represent true power at the desired signal frequency, not the sum of the power contained in all the transverter's mixer products!

It is wise to leave the bandpass filter in the system to assure clean output. If, as is usually the case, your LO is *lower* than the desired mixer product on 1296 MHz, simply back out C4, (tuning toward minimum capacitance or higher frequency) while watching the output of the transverter for a peak. Once this peak is found, tune all screws for maximum output power. It *may require a fair amount of time and patience* to get the feel for the effects of each adjustment, and may even require repeating the entire procedure with a



**The plate tuning capacitor, C4, is mounted under the plate line between the two Teflon pillars.**



Construction of the home-made plate tuning capacitor, C4.

number of available tubes — some fly and some don't!

Once tuneup under reduced power conditions has been accomplished, full B+ may be applied; values from 500 to 1000 volts have been used successfully, but with lots of forced air cooling! The tube must then be re-biased for an idling plate current of 10 to 20 mA *with* LO but *no* i-f signal applied. Then increase CW i-f drive power up to the point of saturation (transverter output no longer increases with increasing amounts of i-f drive) and tune all adjustments for maximum output power. In the ssb mode, talk the transverter output up to an average of about half the maximum CW power output. Typical stage operating parameters are shown in **table 2**. The low efficiency is typical for this type of mixer circuit.

The half-wave, bandpass filter can also be used to transform a 50-ohm transmitter output impedance to 75 ohms to match the impedance of inexpensive,

table 2. Typical operating parameters for the 1296-MHz transverter. Output power was measured at the output of the bandpass filter with a calibrated Bird wattmeter and 50E slug.

| | |
|---|---|
| Local-oscillator power | 3 watts carrier |
| I-f power | 4 watts CW |
| Plate voltage | +800 Vdc |
| Plate current | 150 mA |
| Rf power input | 120 watts |
| Grid bias | −35 Vdc |
| Grid current | 50 mA |
| Rf power output | 17 watts |
| Efficiency | 14% |

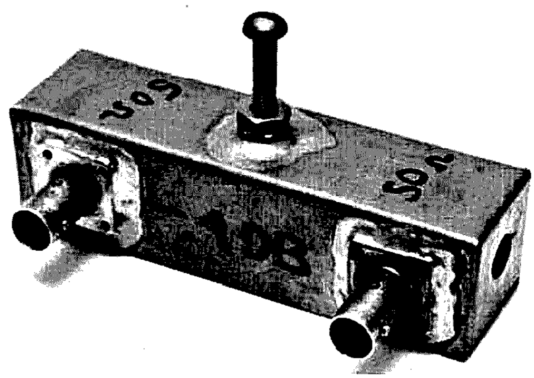
low-loss CATV coax. For 75 ohms simply space the tap 5/8 inch (16mm) from the line end.

## conclusion

It is the intent of this article to describe a relatively simple and inexpensive method for the average amateur to get on 1296 MHz with enough power to make himself heard. Hopefully the ease with which this system can be put together will encourage more activity on the band and finally put to rest the myth that "you can't get there from here on 1296!" I am waiting now to hear from some hard-working and dedicated uhf buffs in Hawaii who would be interested in destroying the current 1296-MHz terrestrial record once and for all! How about it out there, any takers?

## acknowledgements

Sincerest thanks are in order to the many people who contributed to this effort: notably to Bill Jungwirth, WA6NRV, for the beautiful construction of the first working prototype model of the transverter; to Tom Staller, WB6QHF, for the contribution of a commercial varactor tripler; and to the



Half-wavelength transmission-line filter for 1296 MHz. Loss of the filter shown here is 0.4 dB. Dimensions are shown in fig. 8.

West Coast's "Father of 1296," Bill Troetschel, K6UQH, who provided much good advice and encouragement amongst many good ribbings! Photo credits go to Alan Monie. Special thanks go to my wife, Jean, for typing the manuscript.

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**ham radio**

# improved CW transceiver for 40 and 80

Update of a circuit previously presented in *ham radio* — a complete break-in and electronic-keyer circuit have been added to make a truly effective station for traffic and contest work

**On Field Day 1968** I introduced a homebrew 10 - 75 watt CW transceiver for 80 and 40 meters. It performed so well during the contest that Rich Klinman, W3RJ, immediately adopted it, improved it, and used it as a home station. Since then it has been in almost constant use and has survived many a rigorous adventure, finally winding up as my home station.

An article describing the transceiver[1], noted that the lack of break-in operation was one of its more annoying shortcomings. That deficiency has now been corrected, and an ultra simple, built-in electronic keyer has been added to make the unit a truly complete CW station.

This article describes the break-in system and keyer, both of which have some novel features. The objective is to stress the applicability of each to other homebrew or commercial equipment. It's not necessary to duplicate the original rig to use the ideas presented here. A casual inspection of the original article reveals numerous areas in which present technology could be used to simplify the hardware, but we're concerned here only with the functional improvements that have been made.

Many forms of T/R switching are passed off as "break-in;" perhaps the most common being those that switch the transmitter on when the key is depressed and transfer control to the receiver after a delay of an appreciable fraction of a second. These systems drop out upon a deliberate pause but not during normal sending, whereas a really useful system permits interruption by the receiving operator at any time. The circuit described here is of the latter type, in which full recovery occurs even between dots at normal sending speeds.

To describe the break-in system, it's necessary to review a few details of the rig, which is a simple affair using a semiconductor, direct-conversion receiver with phase-shift sideband cancellation for single-signal reception. A 160-meter vfo with plug-in frequency multipliers for 80 and 40 meters provides excitation for both the receiver mixers and the vacuum-tube transmitter. The transmitter uses a 12AU6 driver and a 1625 final-amplifier tube. Both transmitter stages are cathode keyed through a transistor circuit that also provides sidetone and receiver audio blanking. Adjustable offset between transmit and receive frequencies is provided.

In the original design, the T/R function was performed by a dpdt toggle switch. One pole of the switch transferred the antenna from receiver to transmitter; the other pole supplied dc voltage to the keying circuits and also selected one of two reed relays, which controlled the receive/transmit frequency offset. The receiving reed relay connected a front-panel incremental-tuning capacitor across the vfo tank, while the transmit relay substituted a fixed capacitance equivalent to the midrange value of the incremental control.

All functions performed by the toggle switch must be duplicated at high speed by the key to achieve full break-in. Furthermore, a certain sequence of events

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is necessary. The required sequence for a single dot or dash is:

1. Blank receiver audio.
2. Shift vfo frequency.
3. Transfer antenna to transmitter.
4. Turn on transmitter and sidetone.
5. Turn off transmitter and sidetone.
6. Transfer antenna to receiver.
7. Reshift vfo frequency.
8. Restore receiver audio.

It is of course absolutely essential that antenna transfer and frequency shift be completed before activating the transmitter to avoid receiver damage, keying-waveform distortion, or the introduuction of a chirp. For the same reasons, the transmitter output must be allowed to decay fully before the reverse sequence is initiated. The order of sequences **1**, **2**, **3** and **6**, **7**, **8** is somewhat less critical; it's only necessary that audio blanking be initiated early enough and maintained long enough to block thumps due to front-end transients or received-signal chirps due to the vfo shift. In practice, simultaneous occurrence of the events yields satisfactory results if the transitions are fast.

The most crucial aspect of break-in design lies in the antenna coupling method. Low-loss switching of the transmitter output without relays is difficult, so most break-in designs leave the transmitter permanently connected to the antenna and provide means for disconnecting or limiting the rf voltage to the receiver during transmit. A classical approach is that of a cathode follower with a large grid-leak resistance coupled through a small capacitor directly to the coax antenna feed line. Grid current flow in the follower develops a negative bias that limits rf voltage to the receiver.

A serious disadvantage of feed line coupled receiver pickoff is the phenomenon of suck-out. The transmitter tank, usually a pi network, presents a low impedance to the feed line, thereby attenuating received signals. Such attenuation may be tolerable in receivers with a large gain and noise-figure margin but not in simple designs with limited sensitivity.

One solution to the suck-out problem involves coupling the receiver to the plate side of the final tank, so that conditions that provide proper transmitter loading also maximize the received signal. The cathode follower may be so coupled if a capacitive divider is used to reduce the rf voltage applied to the grid to manageable proportions. An additional advantage of the method is that the transmitter tank provides additional selectivity in the receiver front end, thereby improving rejection of out-of-band signals, which might produce intermodulation effects.

For the CW transceiver, it was desired to use an all-semiconductor T/R circuit in the interest of conserving filament power. It was also essential that no loss of gain be incurred and that the circuit introduce no intermodulation problems. These objectives were met using the hot-side coupling approach in conjunction with a shunt-transistor clamping circuit to protect the emitter-follower receiver pickoff.

**T/R switch.** The T/R circuit is shown in **fig. 1**. A 2N709 emitter follower, biased by network R1, R2, drives the low-impedance coaxial-line input to the receiver. Rf from the final tank is coupled to the 2N709 base through a capacitor constructed from a 4-inch (102mm) length of RG-58/U coaxial cable (capacitance approximately 10 pF). During receive, the 2N2222 shunt clamping transistor is inactivated by grounding the control line.

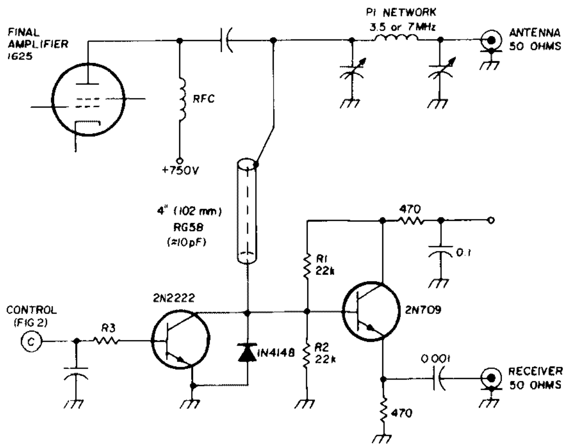



**fig. 1. Solid-state T/R switch. In the original circuit (reference 1) transmit-receive toggle switch. In the improved version shown here true break in is obtained using a 2N2222 and 2N709.**

For transmitting, +12 volts is applied to the control line. The resultant base current through R3 turns on the 2N2222, decreasing the 2N709 base voltage to near ground and cutting off receiver coupling. When the transmitter tank rf voltage goes positive, the 2N2222 conducts more heavily while remaining saturated so that insufficient rf voltage appears at the 2N709 base, which would cause energy transfer to the receiver. On negative rf swings, the 2N2222 conducts to some degree, but the transistor gain under these conditions is poor and assistance is provided by diode CR1, which clamps the negative excursion at about one volt.

The rf current that must be handled by the 2N2222 and the clamp diode is a function of tank rf voltage, frequency, and the value of coupling capacitance. With 750 Vdc on the 1625 plate, the rf voltage approximates 750 V peak, and the resultant peak current through 10 pF is 165 mA at 3.5 MHz and 330 mA at 7 MHz. The 2N2222 base current of about 22 mA is

sufficient to maintain saturation with collector currents of these magnitudes.

An interesting aspect of the circuit is that the shunt transistor and clamp diode need not have spectacular rf response; in fact, a long storage time in the transistor aids to maintain saturation and reduces base drive requirements. During receive, both transistor and diode are cut off and are totally out of the act; therefore they contribute no loading or distortion on high-amplitude signals. The 2N709 follower is degenerative and has good linearity.

It is essential that the control voltage *always* be applied during transmit; a few microseconds at 750 volts will wipe out the whole semiconductor complement! There is a slight difference in optimum tuning points for transmit and receive because the 10-pF coupling capacitor appears as part of the tank capacitance during transmit. The effect is small and may be ignored if tuning is set for optimum transmitter loading. Sensitivity with the T/R circuit is considerably better than that achieved with the receiver

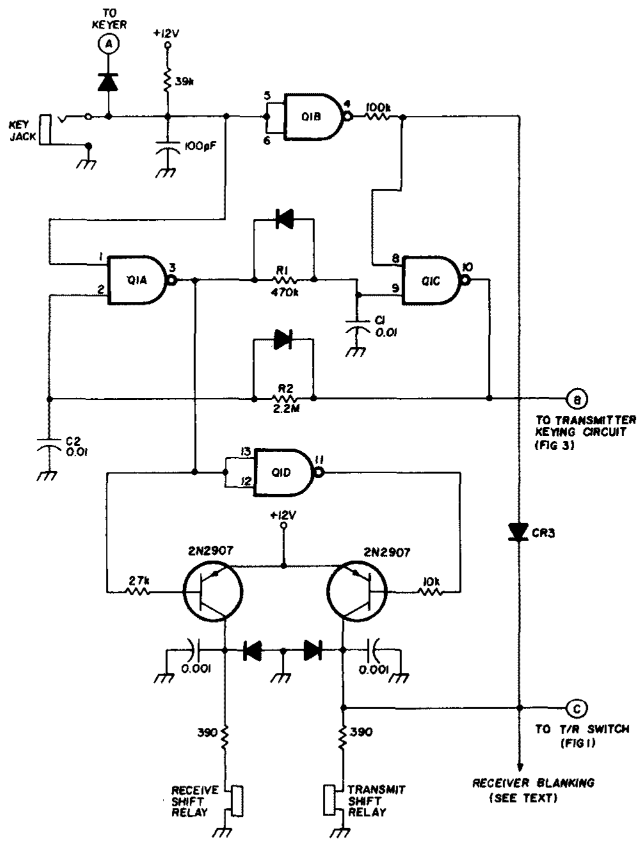



**fig. 2. Sequencer circuit for CW transceiver break-in operation. Connect Q1-14 to +V and Q1-7 to ground. Q1 is a CD4011AE. All diodes are 1N4148. All resistors are in ohms ±10% ¼ watt unless noted otherwise. All capacitors in μF 50 WV unless noted otherwise.**

coupled directly to the feed line (transmitter disconnected), and hash from upband shortwave broadcast monsters is to some degree attenuated.

**Sequencer.** The sequencing portion of the break-in system employs an RCA CD4011AE quad cmos NAND gate in a somewhat unorthodox delay circuit (**fig. 2**). The transmitter keying signal, taken from the output of gate Q1C, is held in the *high* (transmitter off) state until both inputs of Q1C go high. One input, supplied through inverter Q1B, is high when the key is closed; the second input goes high after a delay determined by R1C1. Thus, transmitter turn-on is delayed, but turn-off occurs immedately upon opening the key. On the other hand, the blanking, T/R, and shift functions are controlled by gate Q1A which goes high immediately upon key closure but is held high after key release by voltage fed back through the delay network R2C2. The delay networks include shunt diodes for fast restoration of quiescent conditions. Diode CR3, a redundant safety feature, prevents transmitter turn-on in the event that T/R control voltage is lost.

The rf portion of the original reed relay frequency-shift circuit was left intact. The miniature reed relays (which operate in less than a millisecond) are driven by 2N2907 buffers, with the fourth cmos gate used as an inverter for transmit shift. Receiver audio blanking (not shown) consists of a 2N3565 npn transistor shunted across the volume control.

**Keying circuit.** The transmitter cathode keying circuit (**fig. 3**) requires comment. In the original design, the 12AU6 and 1625 were keyed through separate transistors, with the cathodes unreferenced to any fixed voltage under key-open conditions. Despite the fact that no backwave was present with this circuit, severe interference with receiver operation resulted when the T/R switch was installed. It was necessary to key both stages from a common transistor (40327) and to bias the cathodes to approximately +100 volts under key-open conditions. Positive bias is obtained from network R3, R4, and R5 which also supplies the driver screen; the excitation-control pot, which was always set at maximum, was eliminated. Finally, the electrolytic keying-lag capacitors, which were of insufficient voltage rating, were replaced by a 1 μF, 400-volt paper capacitor.

**Construction.** Layout and construction of the break-in system is noncritical except for the T/R portion. This circuit must be located near the final tank but should be shielded from direct coupling to it. An ideal spot is on the grid side of the final-amplifier partition, with the coaxial coupling capacitor feeding through or looped around the partition. The T/R transistor circuitry should be compact.

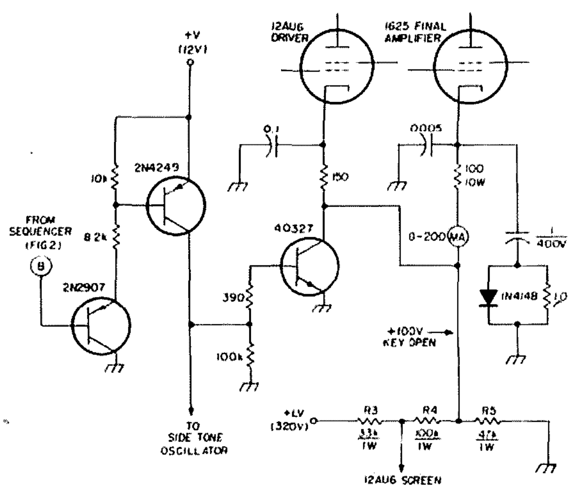



**fig. 3. Transmitter keying circuit for transceiver break in operation. All resistors are in ohms ±10% unless noted otherwise. All capacitors are in μF 200 WV unless noted otherwise.**

**Performance.** Operation is a CW man's delight. Aside from its harmonic content the sidetone signal sounds like any other on the band, and the merest tap on the receiving operator's key brings a screeching halt for "fills." Messages no longer need to be preceded by the frowned-upon "no QSK here," and DX is no longer called unwittingly from beneath somebody else's kilowatt. Contest operation is a breeze.

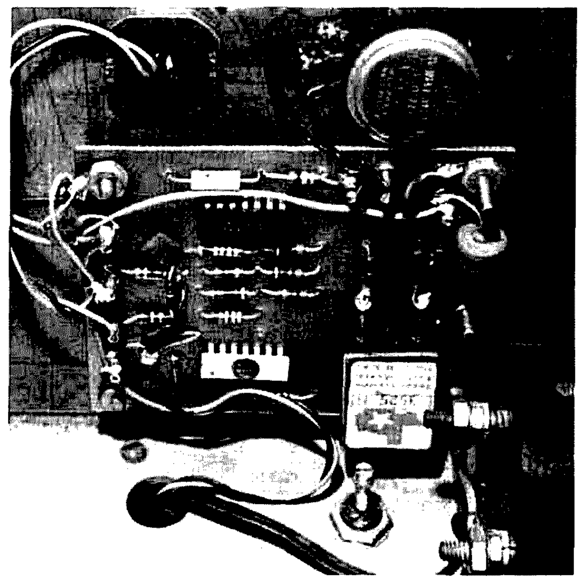
Various improvements will no doubt occur to the reader. For example, in a new design it would be logical to use a varactor diode for incremental tuning and frequency shift; this modification could be made with minor changes in the break-in circuit. Limitations of the system will also be apparent in an all-band transmitter; the clamp current might be excessive at the higher frequencies unless the coupling capacitance is reduced, resulting in loss of sensitivity at the lower frequencies. A compromise value for the capacitance should be possible.

## cmos keyer

The cmos keyer (**fig. 4**) uses two low-cost CD4001AE quad 2-input NOR gates. Two of the gates (Q1A and Q1B) are used as the time-base multivibrator, and the remaining gates (Q1C and Q1D) form a dash flip-flop. Three of the remaining gates are used to synthesize a three-input NOR gate as required for dash control; the last of the eight gates controls the time-base multivibrator and also provides keyer output. Keyer output is positive (at the supply voltage) for **space** and at ground for **mark**, as required by the transceiver. Inversion or level-shifting circuitry could be added easily to adapt the circuit to other forms of keying; or a transistor driver could be added for relay keying.

Operation of the circuit is as follows: under quiescent conditions, Q1A is low, Q1B is high, Q1C is low, Q1D is high, Q2C is low, and Q2D is high. If the dot input is momentarily grounded, Q1A goes high, thereby causing Q1B and Q2D to go low, initiating a dot. If the key is released, the action stops after one dot because Q2D goes high again and locks Q1A in the low state; but if the key remains closed, the multivibrator runs free and produces a string of dots.

If the dash input is grounded, a different sequence takes place. One side of flip-flop (Q1C) receives a positive pulse of width determined by R1C1 each time a character is initiated. The effect of this pulse is to keep Q1C in the low state so that it has no effect on the operation of Q2D. However, the dash input feeds both the multivibrator and Q2A; grounding of the input to Q2A allows a positive pulse (of length determined by R2C2) to be fed Q1D at the same time Q1C is being reset. Since time constant R2C2 is longer than R1C1, Q1D conducts and the flip-flop changes state, thereby clamping Q2D in the low con-



**Cmos keyer board mounted on the sidewall near the transceiver front. The Bendix connector to the keyer paddle and speed-control pot are shown above the circuit board. A frequency spotter (not mentioned in text) is also on this board.**

dition. Q2D remains low throughout the first multivibrator dot-space cycle; the flip-flop is reset at the beginning of the second cycle, but Q2D remains low because Q1A is again high. Thus, Q2D remains low for a period equal to dot-space-dot, as required for proper dot-dash ratio.

The connection from Q1C to Q2C prevents spurious setting of the flip-flop on the second

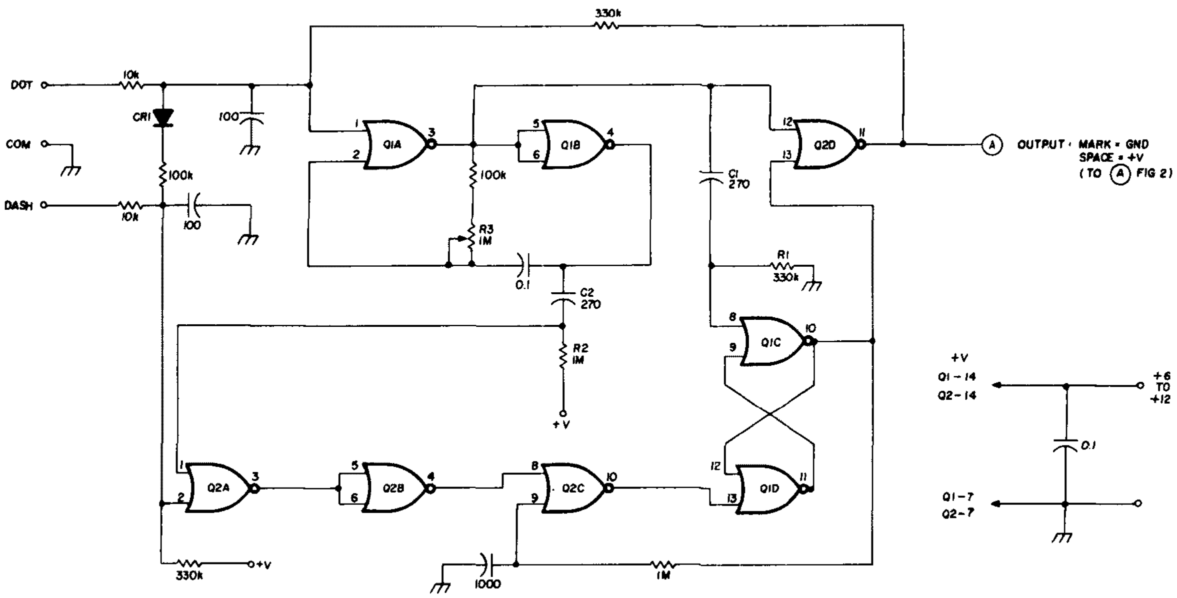



**fig. 4. An ultra simple cmos keyor. The circuit uses two low-cost CD4001AE quad 2-input NOR gates. A triple 3-input NOR gate (CD4025AE) may be substituted for Q2, which allows an output for other functions.**

multivibrator cycle. The 100k resistor in series with gate diode CR1 guarantees dash flip-flop setting at a higher voltage across the key than that required to start the multivibrator, which eliminates a tendency toward production of dots from the dash side; this occurred occasionally with dirty contacts.

Keying weight is determined by the multivibrator duty cycle. This, in turn, is related to the cmos input threshold voltages, which should be half the supply voltage for the desired 50% on time. Since cmos thresholds do vary (from about 33% to 67% of supply), some experimentation may be necessary to determine the best pair of gates for the multivibrator.* Duty cycle can be measured by connecting a high-impedance multimeter or vtvm to the keyer output; the meter will read half the supply voltage on high-speed properly balanced dots.

**Construction.** The keyer was constructed on a piece of glass epoxy circuit board without copper. Holes were drilled for component leads (including the IC pins). The components were inserted and leads crimped over. Point-to-point solder connections were made with small bus wire covered with Teflon sleeving.

This is admittedly a cheap-and-dirty construction but it has survived all kinds of banging around without failure. The circuit board and speed control, R3, are mounted on an inside wall of the transceiver. The key paddle was connected through a 3-wire cable and connector. In the transmitter keying circuit (**fig. 2**), a diode prevents the keyer from interfering with straight-key operation.

**Performance.** The keyer performs beautifully despite its simplicity. It draws no measurable current in the standby mode and only a fraction of a milliamp when running full tilt. The supply voltage is not critical; any voltage from less than 6 to 15 volts can be used. The upper limit is imposed by maximum cmos ratings. The speed, which is adjustable from less than 10 to more than 70 wpm, is nearly independent of the supply voltage. With the key line RC filtering shown, there's no problem due to rf pickup.

The reader with a selection of cmos devices may wish to substitute a triple 3-input NOR gate, type CD4025AE, at Q2 in **fig. 4**. In this case, one of the three gates replaces Q2A, Q2B, and Q2C; the second (with two of its inputs tied together) replaces Q2D, and the third remains free for output inversion or other desired functions. However, if cmos is to be purchased specifically for the keyer, it's better to stay with the two CD4001AEs to provide a wider choice of multivibrator gates.

## acknowledgement


I'd like to express my thanks to Mark Pintavalle, who did the photography.


## reference


1. Clifford Bader, W3NNL, and Richard Klinman, K3OIO, "CW Transceiver for 40 and 80 Meters," *ham radio*, July, 1969, pages 14 - 25.


**ham radio**

* It is also possible to modify the multivibrator circuit to permit adjustment of the duty cycle. See *RCA Application Note ICAN-6267*.

# high-resolution spectrum analyzer for single sideband

All you need is an oscilloscope and the simple circuits described here to display quantitative information about your ssb transmitter

**Most amateurs keep abreast** of recent developments in ssb transmitters and transceivers and, if you're like I am, you wonder how your equipment stacks up against the current crop. Such information, aside from boosting or busting your ego, can help you decide when equipment performance has improved enough to justify purchasing or building a new rig.

One of the most prominently reported performance parameters in equipment-review articles is intermodulation distortion (IMD). Often the IMD performance of an ssb exciter or linear amplifier is displayed as a photo taken from the CRT of an ssb spectrum analyzer. It occurred to me more than once that it would be desirable to be able to produce this kind of photo in my own station. The big hangup has been the lack of a high-resolution ssb spectrum analyzer.

This article describes the technique and hardware to allow the presentation of ssb transmitter IMD products on your oscilloscope. The hardware will cost less than $15 if you purchase everything new, and assembly should take no longer than one or two evenings. No permanent modifications to any of your other equipment will be required.

Also, for anyone not desiring to display the IMD data on a spectrum analyzer, a simpler (no cost) technique is described that will yield the IMD data to be recorded manually.

Most amateurs are familiar with the terms spectrum analyzer and panoramic adapter (Panadapter). Many have such units which have been obtained through military surplus or purchased new. Most of this equipment falls into the panoramic adapter category as defined in a *QST* article on the Heath SB-620 Scanalyzer.[1] Reference 1 differentiates between the two types of instruments, describes the basic theory of operation, and points out the most important performance limitations. That material will not be repeated here, but you're encouraged to consult reference 1 before proceeding further. For those interested in the history of spectrum analysis, more detail concerning analyzer design, or various analyzer applications, references 2, 3, 4, and 5 are recommended.

Since most of the units in the hands of amateurs are low-resolution devices (broad i-f bandwidths), they are not suitable for analysis of narrowband signals such as a two-tone test signal transmitted by an amateur ssb transmitter. Such an analysis requires a spectrum analyzer with a maximum resolution appreciably better than the spacing between the two tones. Laboratory-grade instruments costing thousands of dollars and designed for this purpose of course do the job handily. At present, the only alternative to the amateur is to buy the Heath SB-620 which, from its specifications, should provide the measurement of IMD product levels. The previous Heath Ham-scan (Model HO-13) instrument does not have sufficient resolution to perform these measurements.

The technique and hardware described below will enable the amateur not having a Heath SB-620 or access to a laboratory ssb spectrum analyzer to make scope presentations of transmitter IMD performance much like those seen in the equipment reviews of *QST*.

## theory of operation

The block diagram of a spectrum analyzer is shown in fig. 1. To undertake construction of this whole system would be a major project. However, if you consider this block diagram in terms of equipment already present in many amateur stations, the problem becomes manageable. The blocks in the upper dotted box are part of most amateur receivers. The term *narrowband* i-f is of course relative, and the requirements of this particular application are discussed later in more detail. For the present, let's consider that narrowband means the narrowest selectivity position provided by the typical high-quality receiver or transceiver.

The blocks in the lower dotted box of **fig. 1** are part of all oscilloscopes. Some sweep generators are triggered; some are recurrent. Some vertical amplifiers are ac-

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coupled only; others are dc-coupled. The scope sawtooth may or may not be brought out to the front panel. All these factors will affect the ease with which a particular oscilloscope can be adapted to this application; however, the basics are all there.

The two blocks in **fig. 1** not included in either your receiver or oscilloscope are the swept oscillator and the detector. You may argue that your receiver has a detector and indeed it does. As we shall see, though, it's output is not easily accessible in a form that would be useful for spectrum-analyzer work.

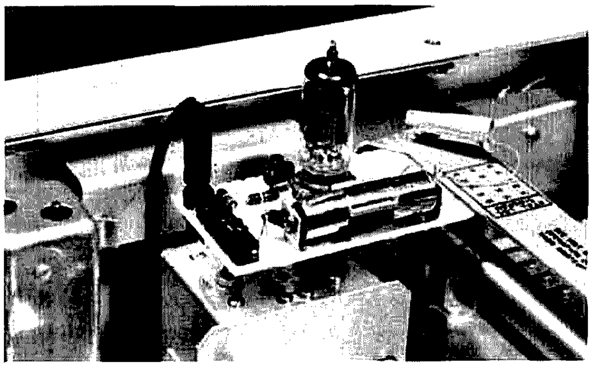
The receiver does have one or more local oscillators. If a way could be found to easily sweep this oscillator or oscillators one of our unaccounted-for blocks would be accommodated. Likewise, a detector can be a simple diode and filter network if we can determine the appropriate place to connect it.

**Swept oscillator.** Since a typical ssb two-tone test will employ tones from 1 to 2 kHz apart, and most applications will require the observation of only third- and fifth-order IMD products, the maximum sweep-frequency range required is 10 kHz. If you're interested in high-order products or wider tone spacing, the sweep frequency range will have to be increased.

Most modern amateur receivers are double-conversion superhets with a crystal-controlled first local oscillator (LO) and a variable-frequency second LO. Typical of this arrangement is the Collins 75S-3. The variable frequency oscillator (vfo) in this receiver is permeability tuned with a vacuum tube as the active device. It tunes from 2.7 to 2.5 kHz to give 200-kHz-wide bands. I decided it would be easier to vary the vfo frequency the required 10 kHz rather than that of the crystal-controlled first LO.

The vfo in the 75S-3 is a Colpitts oscillator; the basic circuit is shown in **fig. 2**. Collins biases diode CR1 on or off to achieve the proper vfo offset frequency when switching sidebands. I decided to use this same point (cathode of the oscillator tube) to apply my variable reactance, thus achieving swept-frequency operation. I elected to use a voltage-variable capacitance diode (varactor) as the variable-reactance source. The varactor tuning diode displays a varying junction capacitance as the reverse bias across a P-N junction is changed. (For more details on varactor diodes, see references 6-9.)

The varactor tuning diode is placed in the circuit of



**Top view of the Collins 75S-3 with sweeper module installed.**

fig. 2 so that the variable capacitance is from the tube cathode to ground. Bias voltage is obtained from a fixed battery source. The oscilloscope sawtooth output gives a repetitive excursion over a precisely set range of voltages that can be selected to yield the desired sweep frequency range.

**Detector.** As discussed previously, the receiver has a built-in detector; it puts out an audio tone near 1 kHz

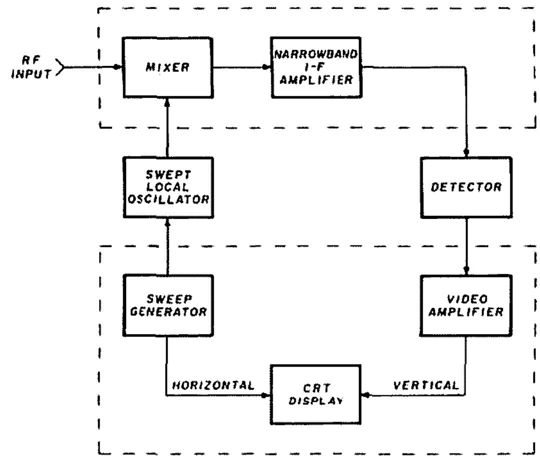



**fig. 1. Basic block diagram of a spectrum analyzer.**

whenever a signal is in the CW filter passband. This signal could be directly displayed on your oscilloscope, but the display would not be what is normally seen on a spectrum analyzer screen. You would see something like the display represented in **fig. 3A** as opposed to the more normal presentation of **fig. 3B**. The waveform filling the pips is the unfiltered 1-kHz audio tone.

For this application, I added an external diode detector (driven by the 1-kHz audio tone), a filter to remove any 1-kHz signal in the output, and a logarithmic shaping circuit to provide the desired dynamic range on the analyzer screen. The block diagram is shown in **fig. 4**. The hardware is built into the black plastic cover of a standard 1/4-inch (6.5mm) phone plug.

## system requirements and limitations

The successful implementation of this system has several prerequisites. First, you must have a receiver or transceiver with adequate selectivity. The selectivity required is typically a 6-dB bandwidth no greater than 1/4 to 1/5 of the frequency to be resolved. That is, for 1-kHz tone spacing, the 6-dB i-f bandwidth should be 200-250 Hz or less. If you use a receiver with a 400 to 500-Hz, 6-dB bandwidth, don't plan to resolve tones much less than 2-kHz apart. Using this rule of thumb, no trouble should occur in resolving IMD products down at least 40 dB from a single tone.

Although I've not experimented with receivers having their narrow selectivity at the audio-output frequency, it should be possible to use them for this application providing care is used to keep signal levels low enough to prevent saturation of any stage preceding the selectivity.

The second prerequisite is that you must disable the agc system in your receiver. If the agc system isn't disabled, low-level signals adjacent to high-level signals on the display will be artificially depressed. If your receiver doesn't have a front-panel switch to turn off the agc, you can determine from the schematic where to break the loop.

Finally, the oscilloscope should have dc-coupled vertical amplifiers with a sensitivity of at least 100 mV/cm. The sawtooth voltage used to sweep the electron beam horizontally across the CRT should be brought out to a terminal on the front panel for use in sweeping the receiver vfo.

Possible system limitations include the lack of an rf gain control on the receiver; this control is desirable to limit the signal level to a point where all stages in the receiver operate linearly. (This function can be performed by a variable attenuator preceding the receiver.) Also, when measuring the IMD of a transmitter located close to your receiver, you must watch for chassis-to-chassis leakage above acceptable limits. This will be a problem particularly with transceivers if you get past the initial problem of operating the receiver and transmitter portions simultaneously. I strongly recommend that in all tests conducted with this technique, provision be made to place the test transmitter at some distance from the analyzer receiver, with the transmitter output attenuated and brought to the receiver through coaxial cable.

## circuit details

The circuit is built in two separate modules: the vfo sweeper and the output signal processor. The sweeper circuit is shown in **fig.** 5. The MV-1404 varactor tuning diode is in series with capacitor C1. This series combination provides capacitance from the oscillator-tube cathode to ground, and thus a variable frequency. The varactor diode is biased by the 1.5-volt battery through R2, R3, and R4. The scope sawtooth (typically 100-150 volts) is reduced by the voltage divider composed of R1 in series with R3 and R4. The attenuated sawtooth is then applied to the diode through R2.

R3 adjustment determines the portion of the

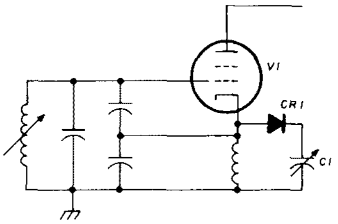



**fig. 2. Basic vfo circuit of the Collins 75S-3.**

capacitance-versus-voltage (C-V) characteristic of the diode over which it operates. In general, it is desirable to adjust the bias and the sawtooth amplitude so that the most linear portion of the C-V curve, which will allow the required 10-kHz sweep range, is being used.

All component values may be varied, as those listed are not the only ones that will provide proper sweep operation. If you use another varactor tuning diode, you'll have to design the circuit values as described above, using the C-V characteristics of the device you choose. For the economy-minded, the least-expensive tuning diode can probably be obtained using the technique described in reference 7.

The output signal processor schematic is shown in **fig.** 6. Diode CR1 rectifies the 1-kHz audio output. The

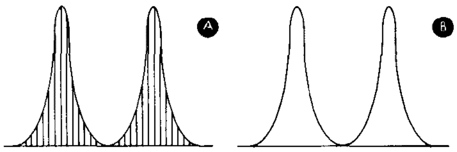



**fig. 3. Spectrum-analyzer display when using a receiver detector, A; at the right is the same display on a more conventional spectrum analyzer.**

signal is filtered by R1, C1. This low-frequency signal is then processed by the logarithmic shaping circuit consisting of R2, R3, CR2, and CR3. Operation of this shaper circuit depends on the fact that the junction voltage developed across the diodes has a characteristic, proportional over a large dynamic range, to the logarithm of the current applied. The 56k-series resistor allows the diodes to operate in a current-driven mode when being driven by the detector output voltage.

## construction

The sweeper module is built on a piece of Vector board with holes on 0.1-inch (2.5mm) centers. The size of the board depends primarily on the size of the bias battery and its holder. Since only six small passive components are mounted on the board in addition to the battery, there's no reason for a large board if you use a battery smaller than the size AA cell shown in the photo. Those with facilities to produce PC boards should be able to lay out and etch this one in short order.

For installation in the Collins 75S-3, I epoxied the board to the top lip of a Pomona Electronics model TVS-7 test socket adapter. You'll have to punch a 5/8-inch (16mm) hole in the board to do this. The oscillator tube can now be removed from the receiver, inserted into the sweeper socket, and the whole assembly can then be plugged into the orignal tube socket. When adapting this scheme to other receivers, take care to select board and battery size that will fit into available space. For receivers that use a solid-state oscillator, it may be desirable to build the sweeper module on the smallest possible piece of board and install it permanently in or near the receiver vfo. No special board layout problems should be encountered, but keep the leads to C1 and CR1 as short as possible. The layout in the photograph is quite satisfactory. All wiring was point-to-point. Components were mounted on one side of the board with wiring on the reverse side.

The output signal processor was built into the handle

of a standard 1/4-inch (6.5mm) phone plug. Use the large plastic handle type, not slim-line or right-angle types. The resistors are 1/4-watt and C1 is a miniature electrolytic. Pigtail leads work quite well with a typical oscilloscope test probe. You may want to affix a

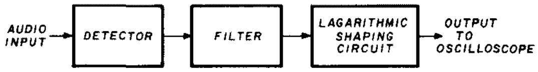



**fig. 4. Block diagram of the output-signal processing system.**

permanent output cable with a connector compatible with your scope; any length cable can be used.

## test and operation

The only alignment required is for the sweeper module sawtooth-level adjustment. If you use any of the Collins S-Line gear and the components specified, the approximate value of R3 (fig. 5) is 3200 ohms. If you're using other equipment with a different vfo frequency, you'll have to use other values for R3, R4, and possibly C1 and the tuning diode. Just follow the general rule of using the most linear portion of the diode C-V characteristic that will allow the required 10-kHz sweep range.

When making your final adjustments to R3, display one of the receiver calibrator markers on the analyzer screen. You should be able to place the marker on the left or rightmost graticule line and, (assuming a 10-cm graticule) by tuning the receiver 10-kHz in the proper direction, move the marker to the other side of the graticule ±1 small division (20mm). A more demanding test of horizontal linearity and sweep range adjustment can be obtained if you have a counter or 1-kHz dial markings on the receiver vfo dial (such as on the 75S-3). With the marker positioned at any one of the cm marks across the graticule, you should move the marker horizontally on the display 1 cm for each 1-kHz rotation of the dial. Likewise, the rotation of the dial through 10 kHz should move the marker across the entire 10-cm graticule ±1 small division. Adjust R3 and, if necessary, the bias voltage to achieve this.

No adjustment is made of the signal processor output amplitude. This is a function of the receiver rf and af gain control settings, rf input signal level, and oscilloscope vertical amplifier sensitivity. These controls are set at the beginning of a series of measurements.

A typical spectrum analyzer test setup is shown in fig. 7. A two-tone oscillator is required to drive the audio input of the test transmitter. Several good articles on the subject have been published, including those by Hank Olsen[10] and Ray Colvin.[11]

Tune the transmitter for desired test power level and terminate the transmitter into a high-power attenuator or a dummy load with an rf coupler to sample the rf for the spectrum analyzer. Locate the test transmitter away from the spectrum-analyzer receiver to prevent unwanted coupling. The *power attenuator* output or sampled rf should be fed to the receiver through a low-power step attenuator.

Remove the receiver vfo tube, insert it into the sweeper module, and install the sweeper in the vfo tube socket. Connect a lead from your oscilloscope sawtooth output to the sweeper input terminal. Be sure to ground the sweeper to the receiver chassis ground. Insert the output signal processor into the phone jack and connect its leads to the oscilloscope vertical input. Be sure the receiver and oscilloscope have a common chassis ground.

You are now ready to display the two-tone test signal with its associated IMD products. Adjust the vertical amplifier sensitivity to approximately 100 mV/cm. Adjust the scope controls for the clearest trace possible. Key the transmitter and adjust the receiver tuned frequency near the transmitter output frequency. You'll find that the receiver calibration will be low by 10 to 15 kHz because of the minimum capacitance added to the vfo circuit by the sweeper module. It may be easiest to tune the receiver with the audio output coupled to your speaker.

After you hear the two-tone signal being swept, you can then plug the output signal processor into the phone jack. Adjust the receiver rf gain control and the step attenuator for the equivalent of an S9 to S9+20 dB signal. Be sure to turn the receiver agc control *off*! Adjust receiver rf gain, af gain, and oscilloscope vertical sensitivity to get a 40-dB dynamic range on the CRT. This can be accomplished by setting the rf and af gain controls to bring the two-tone test signal just barely out of the noise with the step attenuator increased 40 dB beyond a reference setting. Then, after returning the attenuator to the reference setting, adjust the af gain and scope vertical sensitivity for a full screen display. Repeat this procedure several times, and you should have the controls set for a 40-dB dynamic range.

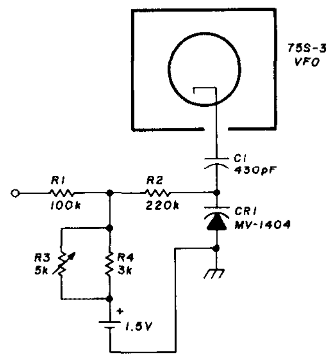



**fig. 5. Sweeper module schematic. All resistors are ¼ watt.**

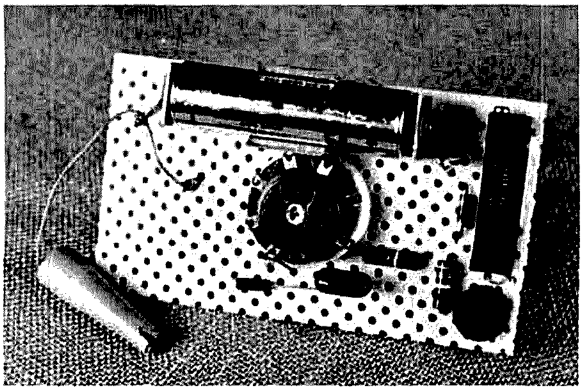
Defining the maximum signal as 0 dB, you'll find that the linear display distance for the first 10-dB increment down is not as great as for the second 10-dB increment, and so on down through the fourth 10-dB increment. This is normal operation and is the price paid for using a relatively simple log shaping circuit. Using a circuit that would give a 10-dB/cm display over the entire 40-dB dynamic range would increase complexity and cost by several times and was not considered worthwhile.

If your step attenuator has 1-dB increments, you can

use it with the 0-dB reference point at the top of the display to calibrate any point to the nearest dB. For most tests, 10-dB calibration marks on the CRT will be adequate. Examples of this type of calibration are shown in the section on results.

### spectrum analysis without hardware

Earlier I stated that a no-cost method of obtaining the IMD performance of a transmitter would be given. The test setup for this method is shown in fig. 8. The same setup as before is used. In this case, the sweep action is accomplished manually by tuning the receiver vfo. The amplitude display is the S-meter. It is not necessary to disable the agc for this method.



**Close up view of the sweeper module showing the small number of parts required. The unit is plugged into the VFO tube socket.**

After the test transmitter is keyed with a two-tone input signal, tune to either of the two tones using the most selective i-f bandwidth available. If you have 100- to 250-Hz CW selectivity, you can resolve tones spaced 1 kHz apart to at least 40 dB. If you have 400- to 500-Hz selectivity, you can resolve tones spaced 2 kHz apart. If your receiver has 1-kHz dial calibration, this test is a lot easier to perform. For instance, if you use 1-kHz tone spacing, the two desired tones will be 1 kHz apart; and third-, fifth- and higher-order IMD products will be spaced at 1-kHz increments above or below the highest and lowest of the two desired tones respectively.

Set the step attenuator and receiver rf gain control so that the S-meter reads about S9+20 dB (this may vary depending on your S-meter characteristics). Just be sure you're not hitting the receiver with so much signal that it is saturating.

Now tune to the desired IMD product and use the step attenuator and rf gain control to set the level of this IMD product to a reference point on the S-meter, such as S-2 or S-3. Keep in mind that you can't increase the signal input to the receiver through the step attenuator to any great extent or the *desired* tones will saturate some portion of the receiver or desensitize it.

Now, retune to the closest desired tone. Increase the step attenuator setting until the desired tone is reduced to the reference level previously set for the IMD product (S-2 or S-3). The increase in attenuator setting is the number of dB that particular IMD product is below a single tone. To obtain the number of dB below PEP output, add 6 dB. Each IMD product (third, fifth, seventh, etc., on each side of the desired tones) must be measured using the above procedure.

This method requires a lot of manual effort but yields the same basic data as does the spectrum analyzer. The advantage of this method is that the only equipment required is a step attenuator and a receiver with good CW selectivity.

### results

The ssb spectrum analyzer, built around the vfo sweeper and output signal processor modules, a Collins 75S-3 receiver with 250-Hz crystal filter, and a Hewlett-Packard model 170A oscilloscope have been in use at my station for about a year. Results have been gratifying. The 250-Hz, 6-dB bandwidth of the 75S-3 CW position allows me to resolve two signals 1 kHz apart to better than 40 dB on the spectrum analyzer screen.

Examples using a signal generator and the 32S-3 transmitter. Fig. 9A shows the spectrum of the SG-823/URM-144 two-tone test signal generator. This test set transmits two tones spaced a variable audio frequency apart. The generator is specified to have all IMD products down greater than 60 dB. This particular display is set up for a dynamic range slightly greater than 40 dB, and no IMD products are visible. This is what the *perfect* SSB transmitter output would look like! In this figure, the two tones are about 2 kHz apart. Ten-dB calibration marks are seen on the vertical centerline and at the side of the picture. The horizontal scale factor is 1 kHz/cm, and the sweep rate is 100 ms/cm.

Fig. 9B is a multiple-exposure photo of the same two-tone signal generator output with the signal level into the spectrum analyzer reduced 10 dB for each successive exposure. Note that the 0 dB reference is on the second horizontal graticule line from the top. The first 10-dB increment is not the same linear distance down on the screen as the succeeding 10-dB increments. A grease pencil was used to mark each 10-dB calibration point on the CRT face; these points were then transferred to the side of the picture.

Fig. 9C shows the spectrum of a Collins 32S-3 transmitter operating at maximum PEP output (117 watts). The frequency is 4 MHz and tone spacing is 1 kHz. The third-order IMD products on either side of the desired two tones are 31 dB down from a single tone (37 dB below PEP, which was determined using the step attenuator). The fifth-order products are barely visible 1

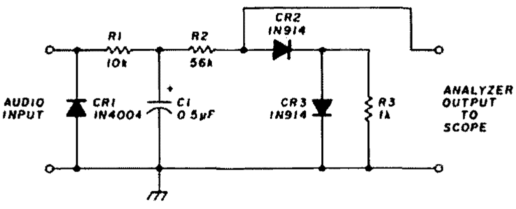



**fig. 6. Output signal-processor schematic. All resistors are ¼ watt.**

kHz on either side of the third-order products; they are just beginning to emerge from the noise. The horizontal scale is 1 kHz/cm and the sweep rate is 100 ms/cm.

**Fig. 9D** is the time-domain equivalent of the 32S-3 output pictured in the frequency domain of fig. **9C**. This is the classic two-tone-test scope picture that amateurs strive to obtain from their ssb transmitters. The picture tells you that the transmitter is performing properly; but the previous picture, in the frequency domain, tells you much more quantitatively and will alert you more quickly that something is deteriorating IMD performance. This picture has a horizontal scale factor of 1 ms/cm. The frequencies of the two test tones are 1 kHz and 2 kHz. Note the 10-dB calibration marks on the vertical centerline used for the spectrum-analyzer display.

**Example using an exciter-amplifier combination. Fig. 10A** is the spectrum of a homebrew linear amplifier driven by the 32S-3 and operating well within its 500-watt PEP output ratings. Tone spacing is 1 kHz. Only the third-order IMD products are visible, which are 35 dB down (verified using the step attenuator). As before, the horizontal scale is 1 kHz/cm and sweep rate is 100 ms/cm.

**Fig. 10B** is a time-domain photo taken under the same transmitter conditions as for **fig. 10A**. As in **figs. 9C** and **9D**, the frequency-domain photo gives much more quantitative information. The horizontal scale for **fig. 10B** is 1 ms/cm and the vertical scale is 200 mV/cm.

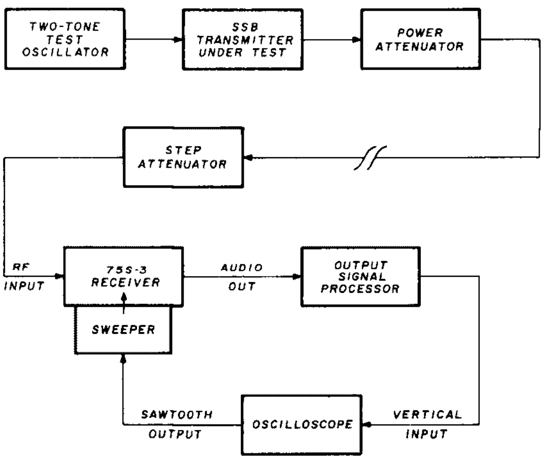



**fig. 7. Typical test setup for measuring intermodulation distortion of an ssb transmitter.**

**Fig. 10C** is of the same linear amplifier-exciter combination used for **fig. 10A**. In this case, however, the equipment was purposely misadjusted. Note that the third-order IMD products are now only 19 dB down. The fifth-order products have increased and are 33 dB below a single tone. The seventh-order products are just barely visible at the noise level. No final amplifier grid current was flowing under the operating conditions but 15 mA (positive) more screen current than normal is flowing. Whenever you see a two-tone test spectrum

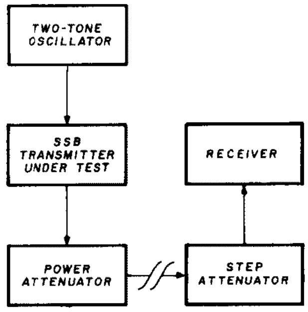



**fig. 8. Test setup for a no-cost intermodulation distortion measurement method.**

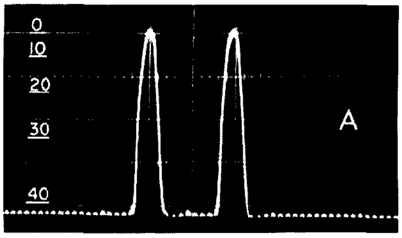
proportioned like this, you can be sure something is defective or out of adjustment. The tone spacing used here was 1 kHz, horizontal scale is 1 kHz/cm, and the sweep rate 100 ms/cm.

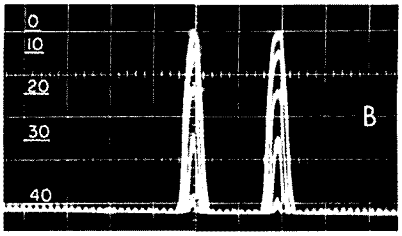
**Fig. 10D** is a picture taken under the same conditions as those of **fig. 10C**. The difference is that the display has been enlarged to fill the picture horizontally. The effective sweep speed in **fig. 10D** is 50 ms/cm; however, in this mode of operation the oscilloscope horizontal amplifier gain was increased by a factor of two – thus no loss of resolution occurred. The horizontal scale is 0.5 kHz/cm, and the vertical-scale 10 dB marks are as before. If I had left the sweep magnification at *X1* (normal) and simply changed the sweep speed control to 50 ms/cm, the display would have occupied the same position as in fig. **10C** but resolution would have been lost (i.e., the pips would have overlapped). Note the ripple on the pips. This is a result of a small portion of the 1-kHz receiver detector audio tone passing through the lowpass filter in the output signal processor. I prefer the appearance of the display in fig. **10C**; **10D** looks a little busy plus the fact that with this expanded horizontal scale you can't see IMD products beyond the fifth-order.

**Examples with equipment misadjusted. Fig. 11A** is of the same transmitter as **figs. 10A** and **C**. In this case, the amplifier was misadjusted and the alc loop disabled. Approximately 25 mA (positive) more screen current than normal was flowing and 0.2 mA of grid current was flowing. The results are degradation of the third-order products to 16 dB below a single tone. The fifth-order products are about the same as those in **fig. 10C**, but the seventh- and ninth-order products are approaching the −40 dB level. The horizontal scale factor is 1-kHz/cm; sweep rate is 100 ms/cm.

**Fig. 11B** shows the time-domain display corresponding to the conditions of fig. **11A**. It is obvious from this photo that something is wrong with transmitter adjustment. I believe you will agree, however, that there is not as much difference between **figs. 9D** or **10B** and this one as there is between **9C** or **10A** and **11A**! The horizontal scale factor is 1 ms/cm and the vertical scale is 200 mV/cm. The two tones are 1 and 2 kHz.

**Example using a phasing-type exciter. Fig. 11C** is the



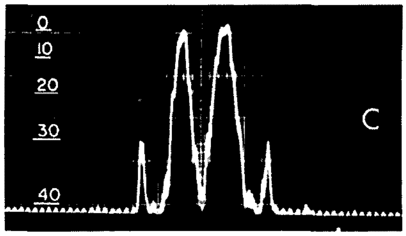
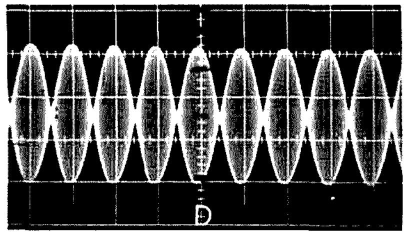




fig. 9. Examples using a two-tone test signal generator and the Collins 32S-3 transmitter.
A. Spectrum of an SG-823/URM-144 two-tone signal generator. Tones are about 2 kHz apart. Horizontal scale: 1 kHz/cm; sweep rate: 100 ms/cm.
B. Multiple-exposure photo of the SG-823/URM-144 two-tone output with signal level into the spectrum analyzer reduced 10 dB for each exposure. Note relationship between each 10-dB increment on vertical scale, which is nonlinear. Horizontal scale factors are as in photo A.
C. Spectrum of a Collins 32S-3 transmitter operating at maximum PEP output (117 watts). Fifth-order products are barely visible 1 kHz on either side of the third-order products. Horizontal scale factors are as in photo A.
D. Time-domain equivalent of the 32S-3 output shown in C. This is the classic two-tone-test scope picture of amateur ssb transmitters. Proper transmitter operation is depicted, but the picture in C gives quantitative information on intermodulation distortion. The two-tone-test frequencies are 1 kHz and 2 kHz. Horizontal scale: 1 ms/cm.



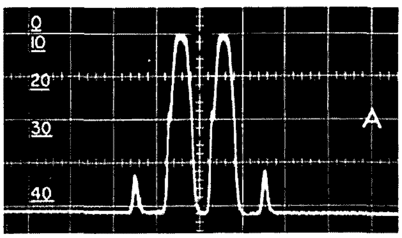

spectrum of a phasing-type ssb exciter. The suppressed-carrier frequency is at the vertical centerline. The two largest signals to the right of the centerline are the desired lower sideband tones. The two pips located 1cm on either side of these two tones are the desired sideband third-order IMD products (26 dB below a single tone). Note that the left (high side) IMD product is at the carrier frequency. The carrier suppression for this rig was checked and found to be greater than 40 dB. So, this pip is the IMD product and not the carrier.

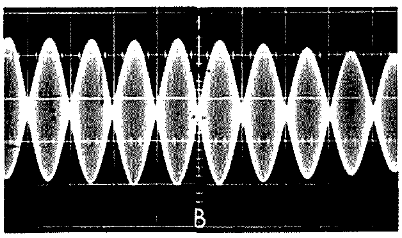
The fifth- and seventh-order IMD products (low side) are visible on the right. Both are about 37 dB below a single tone. At the first- and second-centimeter lines left of the vertical centerline are the two tones of the suppressed or undesired upper sideband. Note that the upper sideband is only suppressed 17 dB with respect to the lower sideband. The third-order IMD product to the right of these undesired pips is buried under the left-side third-order product from the desired sideband. The small pip at the left side of the display could be the high-side third-order product for the two tones of the undesired sideband, or it could be the ninth-order product (high side) of the desired sideband. As you can see, a lot of information is contained in this photo. The horizontal scale factor is 1 kHz/cm and the sweep rate is 100 ms/cm.

## concluding comments

In comparing the time- and frequency-domain pictures here, it's clear that the information contained in the frequency-domain pictures is much greater. Also, when testing or adjusting an ssb transmitter or exciter using the spectrum analyzer, more noticeable changes occur in the display as you change transmitter operating conditions. This is particularly true as you approach optimum transmitter adjustment. That last little improvement is hard to detect on a time-domain picture.

*Some points to ponder in using this equipment:* The



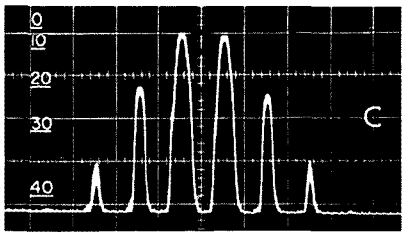
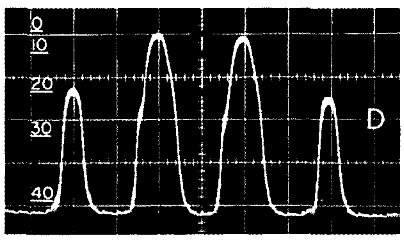




fig. 10. Example using the 32S-3 transmitter and a homebrew linear.
A. Spectrum of a homebrew linear amplifier driven by a 32S-3 operating within its output rating. Tone spacing is 1 kHz. Only third-order intermodulation distortion products are visible, which are 35 dB down. Horizontal scale: 1 kHz/cm; sweep rate: 100 ms/cm.
B. Time-domain spectrum of the setup in photo A. Horizontal scale: 1ms/cm; vertical scale: 200 mV/cm.
C. Same equipment setup as for photo A, except that equipment was purposely misadjusted. Note third-order intermodulation distortion products are 19 dB down; fifth-order products have increased and are 33 dB below a single tone. Seventh-order products are barely visible in the noise. Tone spacing was 1 kHz. Horizontal scale: 1 kHz/cm; sweep rate: 100 ms/cm.
D. Spectrum picture taken under the same conditions as those of C, except that display has been enlarged to occupy entire horizontal area. Scope horizontal amplifier gain was increased by a factor of two to increase resolution. Horizontal scale: 0.5 kHz/cm.



sweeper bias battery life should equal shelf-life of the cell. Most of the pictures were made using a scope sweep speed of 100 ms/cm. This is a relatively slow speed and causes noticeable flicker on a normal oscilloscope CRT. If you attempt to operate at higher sweep rates you'll lose resolution. With 200-Hz receiver selectivity and 2-kHz tone spacing, you may be able to sweep at 50 ms/cm or possibly 20 ms/cm. If you attempt to photograph the analyzer display, the greatest success will be had using single-sweep operation of the scope while holding the camera shutter open (you must use a hood or dark room).

The display on my spectrum analyzer sweeps from left to right, but the highest input frequency is on the left also (increasing frequency from right to left). This condition is brought on by the mixing scheme of the 75S-3. It's quite easy to operate with experience.

The mixing scheme in some receivers will result in more-normal increasing frequency from left to right.

The linearity of this sweeper circuit is quite good considering its simplicity. Referring to figs. 10D and 11C, you can see that the pips are 2 and 1 cm apart, respectivly, ±1 small division. This fact will help you properly identify IMD products by counting centimeter spacing, providing you use tones spaced exactly 1 or 2 kHz apart.

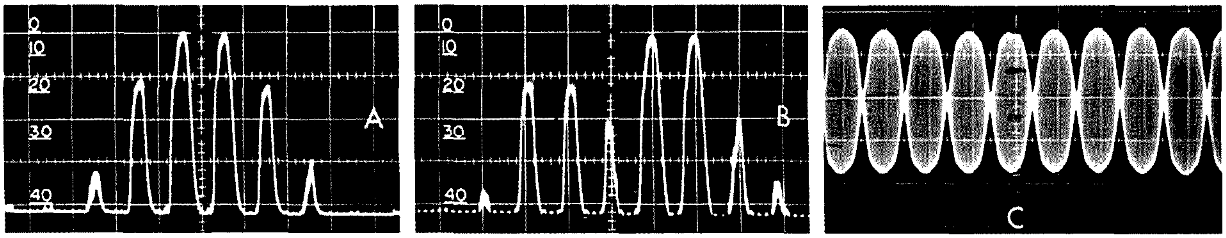
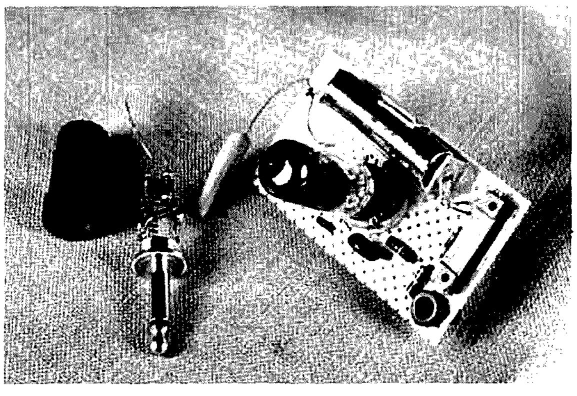



**fig. 11. Examples with equipment misadjusted.**

**A. Spectrum of the same transmitter setup as in figs. 10A and 10C, except that the amplifier was purposely misadjusted and the ALC loop disabled. Third-order products are about the same as in fig. 10C, but seventh- and ninth-order products approach −40 dB. Horizontal scale: 1 kHz/cm; sweep rate: 10 ms/cm.**

**B. Time-domain display corresponding to the conditions of those in fig. 11A. Horizontal scale: 1 ms/cm; vertical scale: 200 mV/cm.**

**C. Spectrum of a phasing-type ssb exciter. Upper sideband is suppressed only 17 dB with respect to lower sideband. Fifth- and seventh-order intermodulation distortion products are 37 dB below a single tone. Horizontal scale: 1 kHz/cm; sweep rate: 100 ms/cm.**



**Details of sweeper and output-signal modules.**

I mentioned earlier that the minimum capacitance of the sweeper module would upset the dial calibration of your receiver. In my case the calibration was offset 12 kHz low. If your measurements require an accurate knowledge of the carrier (suppressed carrier) frequency, I recommend the use of some form of digital dial readout. If you use a readout such as in the Heath SB-650 or the frequency measuring system I described,[12] your readout will automatically account for the sweeper offset. If you use a readout that is preset and adds or subtracts the vfo frequency, you'll have to recalibrate the preset.

I hope this inexpensive method of high-resolution ssb spectrum analysis will encourage more amateurs to reap the advantages of frequency-domain analysis. Once you start using a spectrum analyzer, whether for ssb IMD distortion analysis, harmonic distortion analysis, or in search of oscillations and spurious emissions, you'll find that you'll be quite dependent on the instrument.

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**ham radio**

# rat-race balanced mixer for 1296 MHz

Details for an easily built mixer for the 1296-MHz amateur band which can be used in both receiving and transmitting applications

The heart of most successful uhf ssb transceivers is a bilateral mixer, either singly or doubly-balanced. In previous articles[1,2] I have published designs for several such mixers. All have been built successfully by a number of readers, but each exhibited design characteristics which restricted universal reproducibility. I have thus received numerous requests, primarily from amateurs in Europe and Asia, to develop a mixer which could be built from readily available materials, without specialized dielectrics, advanced metalworking techniques, or expensive commercial microcircuits.

Actually, such a mixer was published some time ago by Paul Wade, WA2ZZF.[3,4] His mixer used a microstripline quadrature hybrid, and can be easily etched on a double-sided, glass-epoxy printed-circuit board at minimal cost. As a receive converter, the unit yields an outstanding noise figure because of its inherently low conversion loss.

Unfortunately, the Wade mixer falls a bit short of the mark in transmit service, due in part to its limited dynamic range, moderate isolation, and difficulties in obtaining a good wideband impedance match. The rat-race mixer described here overcomes some of these limitations and preserves the reproducibility of Wade's design.

## mixer anatomy

Basically, any passive mixer assembly can be functionally divided into three segments, as shown in **fig. 1**. The coupler serves to apply components of the rf and local-oscillator (LO) signals to the mixer diodes in the correct phase relationship, and may consist of transmission-line delay networks, resistive or reactive power dividers, balun transformers, coaxial or waveguide directional couplers, hybrid couplers, or some combination. The coupler may be built with lumped constants, coaxially, with toroidal transformers, in stripline, or in microstrip.

The nonlinear network in which sum and difference frequencies are generated is represented by the diode array of **fig. 1**. Unbalanced mixers (which afford no isolation between the LO and rf ports) use a single diode, while most single-balanced mixers use a matched diode pair. Double-balanced mixers typically include four diodes in either a ring or bridge arrangement, while special-purpose mixers which offer image cancellation or extra-wide dynamic range may contain an array of eight or more diodes.

Diode selection is based upon noise figure, conversion efficiency, and the signal amplitude requirements of the system. Point-contact cartridge diodes, such as the 1N21 series, were the accepted standard in microwave mixers for many years. These devices are still used in high-power radar applications where diode burnout immunity and ease of replacement are major considerations.

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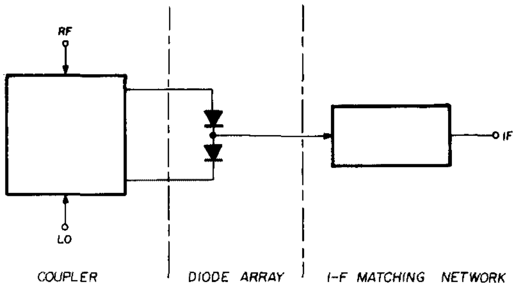



**fig. 1. Basic anatomy of a balanced mixer. The coupler may take several forms as shown in fig. 2. The i-f matching network is designed to match the impedance of the diodes to 50 ohms.**

The current favorite, so far as the uhf experimenter is concerned, is the Schottky barrier or *hot-carrier* diode. These diodes are available in low-cost glass packages from such microwave industry leaders as Aertech, Alpha, Hewlett-Packard, Microwave Associates, Parametric Industries and others, and combine low conversion loss with excellent immunity to rf burnout. An additional advantage, so far as balanced mixers are concerned, is that the manufacturing variations for hot-carrier diodes are minimal. This yields, for any two diodes of the same part number, rf characteristics which are closely matched, thereby eliminating the need for selecting matched pairs or quads of diodes to achieve good mixer balance.

Where the ultimate in conversion efficiency is required (especially at the higher microwave frequencies), tunnel diodes excel. However, their high cost and relative susceptibility to rf overload have limited their acceptance by radio amateurs.

Regardless of the diodes which are selected for the mixer's nonlinear network, it is unlikely that they will provide a good impedance match to 50 ohms at the i-f port. Bear in mind that diode impedance varies with diode current, which in turn is a function of local-oscillator injection level, applied dc bias, and the diodes' dc return path. The complex impedance of the i-f port, under the desired operating conditions, should be transformed to 50 ohms. To minimize intermodulation responses, this transformation should be effective at both the desired i-f frequency *and* the undesired mixing product (usually the *sum* frequency). This will serve to eliminate reflections at any frequency from re-entering the nonlinear network through the mismatched i-f termination. The effects of image frequency termination are discussed in detail in reference 5, as well as section 2.10 of reference 6.

The required impedance transformation at the i-f port can be accomplished by applying conventional techniques to the design of L, T, or pi networks. These networks also serve to suppress any components of the rf or LO signals which may be present at the i-f port because of mixer imbalance.

## hybrid selection

The generalized balanced mixed circuit in **fig. 1** allows considerable latitude in the selection of coupling elements for applying the LO and rf signals to the mixer diodes. Hybrid couplers are available for dividing the input signals into two equal components, which are either in phase, 180° out of phase, or in phase quadrature.[7] Although mixers have been designed around virtually every imaginable coupler, the arrangements considered for this design were two versions of the 90° branch hybrid or quadrature power divider, and two types of 180° ring

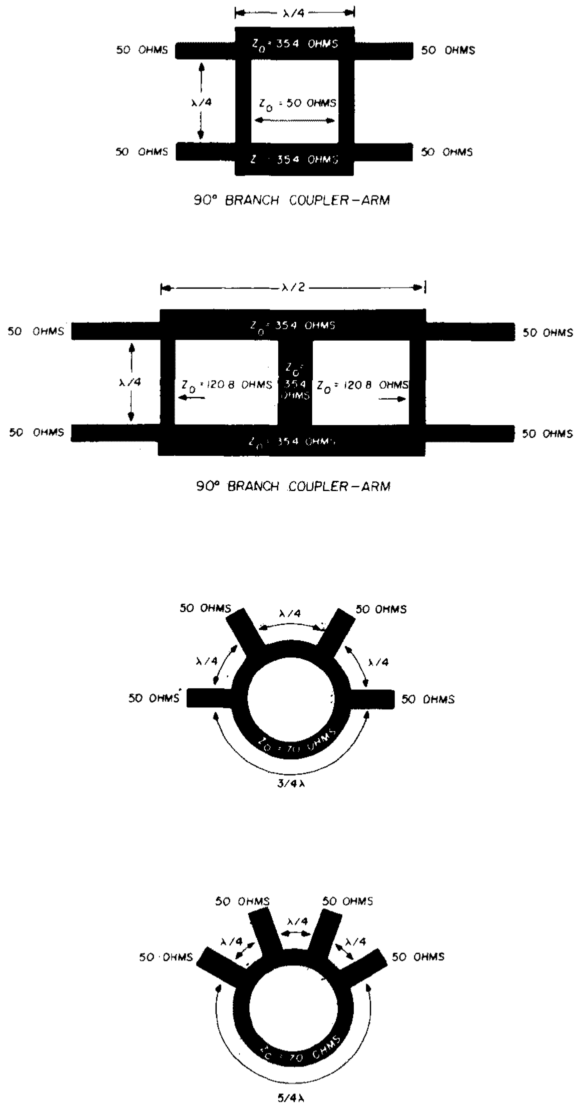



**fig. 2. Four types of microwave hybrid couplers which are used in single-balanced diode mixers. Performance of the various types is compared in table 1.**

hybrid or *ratrace* (see **fig. 2**). Each of these hybrids can be easily built with etched microstripline, and each offers certain performance advantages which should be considered.

Assuming an rf operating frequency of 1296 MHz, a 144-MHz i-f would suggest a LO frequency of 1152 MHz. Since the mixer hybrid must pass both rf and LO components, it is reasonable to specify 1224 MHz, the mean of these two frequencies, as the design center frequency for the hybrid. The ripple passband of the hybrid must, of course, include both the rf and the LO frequencies; thus the design must permit an operating bandwidth of 144 MHz, or 12%. Physical construction of the hybrid in microstripline, however, could introduce several sources of error as to actual center frequency. Because of the effect of variations in substrate dielectric constant, or that of cumulative dimensional tolerances, conservative design philosophy suggests characterizing the hybrid over a somewhat greater bandwidth. For this study a 20% operating bandwidth was assumed.

**Table 1** compares various operating characteristics of the couplers shown in **fig. 2** over this 20% bandwidth. Information regarding vswr, isolation, and imbalance was determined from published design charts;[8] the relative microstripline losses at the center frequency were derived empirically.

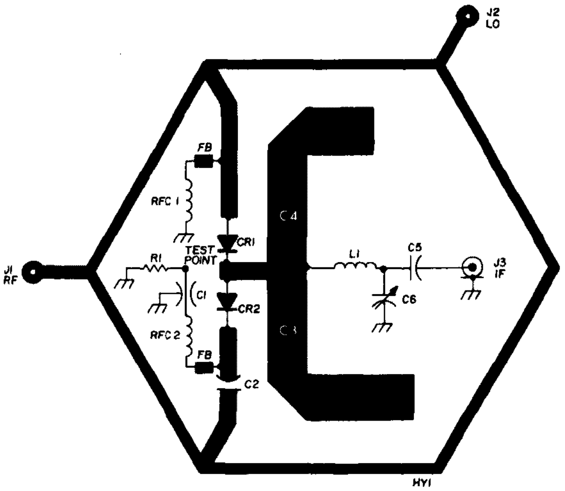
Since all of the listed hybrid characteristics contribute in varying degrees to the overall performance of the resulting mixer, a somewhat subjective tradeoff process is necessary to select the optimum mixer hybrid. I began by excluding the popular 90° branch coupler because of its high vswr and relatively poor isolation. Similarly, I eliminated the extended ring hybrid on the basis of its high stripline losses and poor amplitude balance. In selecting between the two remaining candidates, I opted for minimizing amplitude imbalance and stripline losses, trading off

**table 1. Performance comparison of mixer hybrids over a 20% bandwidth.**

| hybrid type | vswr | amplitude imbalance | isolation | relative insertion loss |
|---|---|---|---|---|
| 90° branch – 2 arm | 1.45 | 0.7 dB | 14.0 dB | 1.0 |
| 90° branch – 3 arm | 1.12 | 0.5 dB | 25.3 dB | 1.7 |
| 180° ring – 3/2λ | 1.14 | 0.4 dB | 23.0 dB | 1.5 |
| 180° ring – extended | 1.40 | 0.9 dB | 23.0 dB | 2.0 |

the resulting slight degradation in vswr and isolation. The form factor of the final mixer was a further consideration, although I must admit that a toss of a coin may have been just as scientific!

It is interesting to note that other uhf experimenters have also selected the 3/2 wavelength, 180° ring hybrid for mixer service.[9,10,11] All have achieved respectable mixer performance.




| | |
|---|---|
| C1 | 1000 pF feedthrough capacitor |
| C2 | 100 pF chip capacitor (ATC 100B or equivalent) |
| C3 | 25-ohm open-circuit microstripline, 1 quarter-wavelength long at 1152 MHz (see fig. 4) |
| C4 | 25-ohm open-circuited microstripline, 1 quarter-wavelength long at 1296 MHz (see fig. 4) |
| C5 | 0.001 μF ceramic capacitor |
| C6 | 10-40 pF cermic trimmer capacitor |
| J1, J2 | SMA coaxial receptacle (E.F. Johnson 142-0298-001) |
| J3 | BNC coaxial receptacle (UG-1094/U) |
| L1 | 4 turns no. 20 (0.8mm), air wound, 3/8" (9.5mm) diameter, 1/2" (13mm) long |
| R1 | 10 ohm, ¼ watt carbon composition resistor |
| RFC1, RFC2 | miniature 0.33 μH molded rf choke with ferrite bead slipped over hot end |
| Z1 | 70-ohm rat-race hybrid at 1224 MHz (see fig. 4) |
| CR1, CR2 | hot-carrier diodes (Hewlett-Packard HP 5082-2817) |

**fig. 3. Schematic diagram of the rat-race balanced mixer for 1296 MHz. Full-size printed-circuit layout for the etched circuit board is shown in fig. 4.**

The decision to use microstripline construction on fiberglass-epoxy PC board was due primarily to the success which Wade achieved with this medium, and partly due to my frustration in trying to duplicate the results of others. Rat-race mixers of slab-type[9] or coaxial[10] construction demand greater patience and mechanical ability than I possess, and the slight, almost immeasurable performance advantage of etching microstripline couplers on expensive Teflon substrates[11] didn't seem worth the effort.

## i-f selection

Factors governing the selection of a conversion system's intermediate frequency include the band width capabilities of hybrid junctions, image rejection limitations in practical bandpass filters, the availability of i-f equipment or components, and spurious responses (related to possible harmonic relationships

between signal, LO, and i-f). Unfortunately, the first of these considerations seems to suggest selection of a low i-f, the second generally favors a high i-f, and the third all too often dictates selection of an i-f which violates the fourth!

I have built various conversion systems for 1296 MHz, using amateur 28, 50, 144, and 432 MHz bands as a receive and transmit i-f. My experiences tend to favor the 2-meter i-f as a good compromise between mixer bandwidth and image rejection. The recent proliferation of 2-meter ssb transceivers further supports this choice. However, the actual operating frequency within the band must not be selected arbitrarily. For example, placing a 1296.0-MHz signal at a 144.0-MHz i-f would require a 1152.0-MHz LO. More than one experimenter (myself included) has started with a 48.0-MHz crystal oscillator, multiplied by 24 to 1152 MHz, and ended up with a strong interfering signal at 144 MHz on the receiving dial.

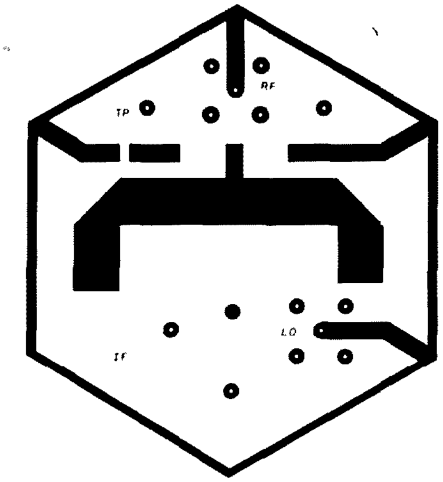


**fig. 4. Full-size layout for the rat-race mixer circuit board. Drilling instructions are shown in fig. 5.**

The spurious signal, of course, is the third harmonic of the crystal oscillator. If you try to circumvent this difficulty by starting at 96.0 MHz and multiply by 12, it's possible to end up with transmitted spurious signals from the ninth harmonic of the injected i-f signal.

Another constraint on the i-f selection is the undesirability of having a popular operating frequency in the 1296-MHz band fall at (or worse, just past) the end of the i-f tuning range. For best results, 1296.0 MHz should fall in the middle of the tuning dial. In my present system, this dictates an i-f of 145.1 MHz; a 1150.9-MHz LO derived from a 95.9083-MHz crystal keeps me free from birdies.

Although most of my activity on 1296 MHz is on ssb, the beauty of a linear translation scheme is that it's compatible with virtually any mode. This mixer will accommodate fm, a-m, video, or whatever. Depending on the mode, tradition, rather than technology, may govern the selection of operating and i-f frequencies. A group planning narrow-band fm simplex, for example, might choose to place 1290.52 MHz on 146.52 MHz. Nonetheless, the above considerations deserve attention, especially when selecting the LO multiplying scheme.

The mixer design presented here will actually accommodate an i-f between about 70 and 150 MHz, with little noticeable performance variation. Obviously, great care should be exercised when selecting the actual i-f and oscillator frequencies.

## construction

Fig. 3 shows a schematic diagram of the ring hybrid or rat-race mixer for the 1296-MHz amateur band. Note that the microstripline hybrid, which is normally built as a circular trace with a circumference 3/2 wavelengths long, is shown as a hexagon with quarter-wavelength sides. I used this same simplification in an earlier mixer,[12] as did Dick Bingham, WB6BDR.[11] Neither of us noticed any measurable degradation over the usual circular arrangement. Leroy May's successful rat-race mixer (reference 9) used a triangular layout, and this led me to believe that the layout itself is relatively unimportant. The 70-ohm characteristic impedance of the ring, however, is critical, and requires a 0.05 inch (1.3mm) microstripline width for G-10 circuit board.

Aside from the diode bias path (to be discussed

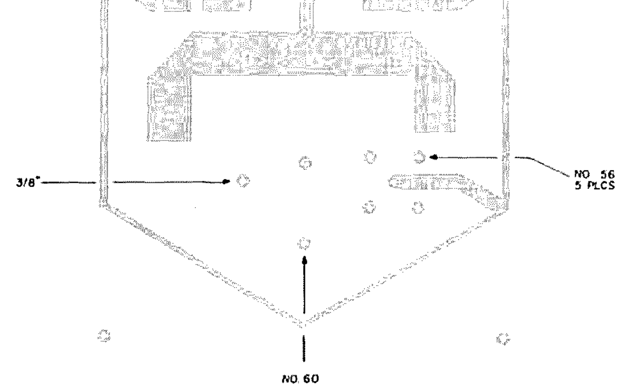


**fig. 5. Drilling template for printed circuit board.**

shortly), the bulk of this mixer design, including the i-f port matching network, is borrowed directly from Paul Wade's design[4] and will not be discussed here. I should point out, however, that the pi-network matching network at the i-f port does little to properly terminate the image component; the importance of this was discussed previously. For mixer applications which dictate maximum freedom from intermodulation distortion, the user should consider adapting the Bridge-T interstage isolator circuit.[14]

**Fig. 4** is a full-scale layout for the rat-race mixer. The pattern should be etched on 1/16 inch (1.5mm) thick G-10 fiberglass-epoxy printed circuit laminate, double-clad with 1 ounce copper (1.4 mils or 36 microns thick). One side of the board is unetched and serves as a groundplane for the microstriplines on the etched side. Drill the board as shown in **fig. 5**.

Assemble the mixer as shown in **fig. 3** and the photographs. When mounting the SMA connectors at the LO and rf ports, be sure to countersink the groundplane side of the board at the connector center conductor to avoid a short to ground. Assembling the mixer is straightforward, except that the leads of the diodes must be bent carefully to prevent damage to the glass diode package.

Once assembled, only one mixer adjustment is required: matching to the i-f port. This can be done by injecting a signal into the rf port, connecting the i-f port to a 2-meter receiver, and tuning capacitor C6 for maximum signal level. Alternatively, with the rf port driving a power meter, inject a 3mW, 2-meter signal into the i-f port, and tune C6 for maximum output. In either case, the LO port must be driven with a clean 5 to 10 mW signal in the vicinity of 1152 MHz and the i-f port must be terminated in 50 ohms. The 50-ohm load can be provided by tuning the mixer with a fixed, 50-ohm pad inserted between the i-f port and the 2-meter transmitter or receiver.

**Fig. 6** shows the isolation of this mixer, as measured on a Hewlett-Packard 8507A Automatic Network Analyzer. The isolation of the assembly is greater than 20 dB from 1050 MHz to over 1300 MHz, and at least 30 dB over the anticipated operating bandwidth; isolation over a 20% bandwidth is quite close to that predicted in **table 1**.

## bias considerations

Wade operates his balanced mixers in the *starved-LO* mode.[3,4] That is, he applies a rather low level of LO injection (typically 1 mW) to the diodes, and supplements this low LO drive with dc bias. By varying the dc bias, the impedance of the diodes can be controlled, providing an improved match to the mixer hybrid. Obviously, it is far easier to generate the required dc bias than it is to increase LO injection by 10 dB.

Other advantages of DC mixer bias were reported by Pound as far back as 1948:[6] "Accompanying the reduced LO power requirement is a reduction in the reaction of the local-oscillator circuit on the signal circuit in the mixer. The over-all noise figure becomes less dependent upon the amount of incident local-oscillator power at the crystal, because the conversion loss does not increase so rapidly as the local-oscillator drive is decreased. Finally, the i-f conductance of the crystal is less dependent on the amount of local-oscillator drive."

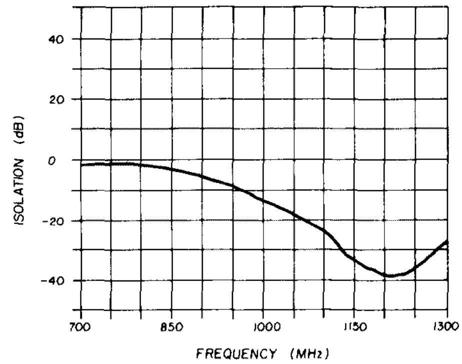



**fig. 6. Isolation of the rat-race balanced mixer, as measured with a Hewlett-Packard 8507A Automatic Network Analyzer.**

*Starved-LO* operation was attempted in one of my bilateral conversion systems, but with disastrous results. In transmit mode application of 1 mW of i-f injection resulted in a severely distorted rf output, measuring several dB below the power level which the characteristic conversion loss of the mixer would suggest. Further, a spectrum analyzer revealed numerous unwanted frequency components which were not previously a problem. It appeared that the dynamic range of a balanced mixer in the *starved-LO* mode was far below that of a similar mixer with the customary +3 dBm (2 mW) per diode injection level. This conclusion was confirmed by Harlan Howe:[8] "Naturally, intermodulation products and saturation levels will be the same as if the LO drive were unsupplemented by dc power." Thus *starved-LO* operation was abandoned for this mixer in favor of a full 4 to 10 mW of LO injection. A similar choice is indicated for any balanced mixer used in transmitter service, or any time large-signal handling capability is important.

It will be noted from the schematic in **fig. 3** that, although external dc bias is not used in this design, the diode's self-bias return is brought out through feedthrough capacitor C1. Since the LO signal is effectively rectified by the diodes, a dc voltage will appear across R1; the magnitude of the dc voltage is a function of applied LO injection level. Thus, the feedthrough capacitor is a convenient test point for

measuring (and optimizing) LO injection, as indicated in **fig. 7**. The object is to minimize conversion loss, which is a direct function of LO injection (see **fig. 8**).

Like other microwave experimenters, in recent years I have sought to master the techniques of microstripline design and construction by developing individual circuits on etched circuit boards. Thus my early 1296-MHz systems consisted of numerous modules, each containing a single stage, each matched to 50 ohms, and all inter-connected by coaxial cable. This approach at best is a crude utilization of microstripline technology. One of the main benefits to be derived from the etched-substrate approach is that numerous associated circuits may be built as an integral unit, thereby minimizing interconnections, reducing cost, and improving reliability.

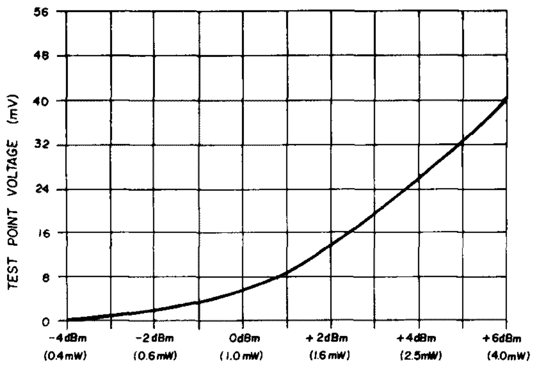



**fig. 7. Since the diodes rectify the LO signal, a dc voltage measurement can be used to check LO injection level. Plotted here is the test-point voltage (mV) vs LO injection for the rat-race balanced mixer.**

I have made several attempts to integrate circuits onto common substrates. My process of system integration began with a two-stage preamplifier,[12] progressed to a combined balanced mixer, image filter, and LO filter,[2] and culminated in a complete transverter-on-a-board.[13]

Unfortunately, acceptance of these integrated assemblies by amateur microwave enthusiasts has not been overwhelming. It was brought to my attention by numerous readers that the tasks of tuneup and testing present major problems when multiple, interactive stages are involved, especially when the available test equipment is limited to a grid-dip oscillator and a number 47 lightbulb! Individual stages, on the other hand, may be readily tested into 50 ohms, and then placed into the desired system with few adjustments. Joe Reisert has been promoting such a modular approach for some time; his articles are highly recommended.[15,16,17]

My 1296-MHz systems have recently gone through a process of de-integration (not to be confused with *dis*integration) with a return to small modules, each containing a single stage, all stages interconnected with 50-ohm coax.[18] This mixer is one such module. Unlike my previous mixer, the one presented here contains no rf or LO filters. Filters will of course be required in most applications but are easily added to the system as individual modules after they have been tuned and tested.[19]

A word about rf cabling and connectors is in order. Whenever possible I recommend the use of high quality microwave connectors such as type SMA. The JCM series of coaxial connectors from E.F. Johnson are low cost, SMA-compatible units exhibiting excellent rf properties through 4 GHz. For microstripline launchers I recommend E.F. Johnson part number 142-0298-001, and I use their 142-0161-001 connectors on all jumper cables. These connectors are priced in the $3.00 range and are available from electronics distributors in many parts of the United States.

Of all the considerations surrounding the selection of interconnecting coaxial cable, cost and availability are the major factors for amateur applications. Obviously, low loss and constant impedance must also be considered. Semi-rigid coax, such as Uniform Tubing UT-141,* perform exceptionally well, but are virtually unobtainable in many parts of the world. A good second choice is ¼ inch (6.5mm) diameter flexible coax, if the lengths are short. Desirable features are double shielding, Teflon dielectric, and a silver-plated, solid center conductor. One cable meeting these requirements is RG-142/U, but its cost per foot discourages many experimenters. Nonetheless, a moderate investment in jumper coax can spare you a world of grief. Other usable cables which cost considerably less include types RG-141/U, RG-223/U, and RG-55/U. As a last resort, RG-58/U may be used, but its length must be kept to an absolute

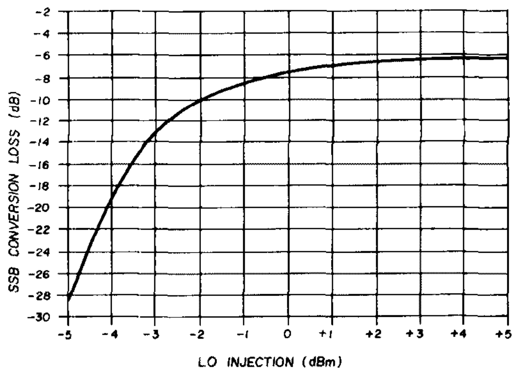



**fig. 8. Conversion loss of the rat-race single-balanced mixer as a function of local-oscillator injection.**

*Not to be confused with RG-141/U, which is a ¼ inch (6.5mm) diameter flexible cable.

minimum. Just remember that most 1296 operators use RG-58/U to build calibrated attenuators! All of the flexible cable types listed here will accept the recommended SMA plug.

## parts availability

The components required to build the rat-race mixer described in this article are available from various sources in the United States; the printed-circuit board may be etched from the full-size artwork shown in **fig. 4**. For those uhf experimenters who don't have the facilities to etch their own boards, commercially etched, drilled, and plated boards are available from Microcomm.*

*An etched, drilled, and plated circuit board for the rat-race mixer is available from Microcomm, 14908 Sandy Lane, San Jose, California 95124, for $6.50 post paid within the United States and Canada ($7.00 elsewhere). A wired and tested mixer is also available; Microcomm will *not* offer a complete kit of parts.

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**ham radio**

# self-supporting coils

## A set of inductance and $Q$ graphs for airwound coils using screws and other threaded forms as mandrels

**Self-supporting coils** without forms are the cheapest, most convenient, high-$Q$ inductors for inductance values under 0.4 microhenries. Unfortunately, the wire is difficult to wind on a smooth form and you have to account for wire diameter plus form diameter for accurate calculation. Presented here is a method for using conventional screw threads as forms along with measured inductance values.

Form and wire size, coded by the letters of **fig. 1**, are given below:

| curve | wire size | winding form | |
|---|---|---|---|
| A | 26 enameled or Soldereze | 4-40 | screw |
| B | 22 enameled or Soldereze | 6-32 | screw |
| C | 22 enameled or Soldereze | 8-32 | screw |
| D | 22 enameled or Soldereze | 10-32 | screw |
| E | 18 tinned or enameled | 1/4-20 | screw |
| F | 18 tinned or enameled | 5/16-18 | bolt |
| G | 18 tinned or enameled | 3/8-16 | lag bolt |
| H | 14 bare copper | 7-watt Christmas bulb base (3/8″ ID, 10 TPI) | |
| J | 12 bare copper | Paint roller ferrule (5/8″ ID, 1 turn per 3/16″) | |
| K | 12 bare copper | Standard 117 Vac lamp base | |

The method is very simple: Take any convenient screw or bolt and wind the wire firmly on the threads. When finished, allow the wire to release its tension, then carefully remove the form. **Fig. 1** shows inductance vs the number of turns for ten different screw threads on easily obtained sizes.

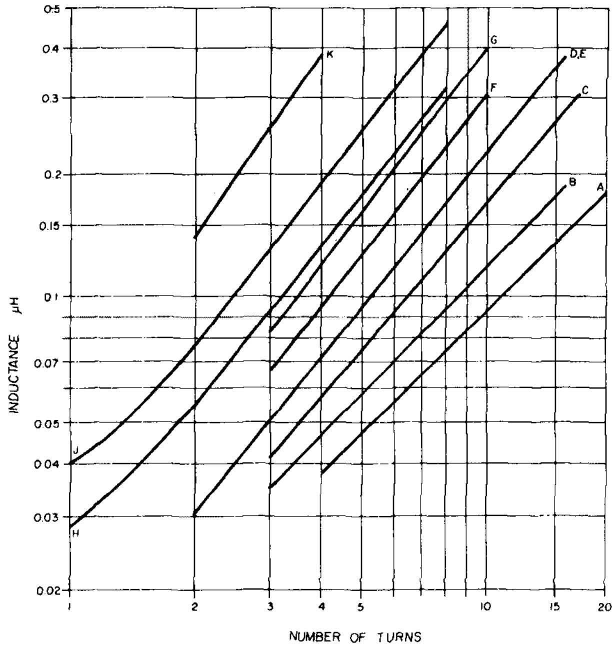


**fig. 1. Inductance of airwound coils using threaded forms as mandrels. Extensive use of this chart has shown the inductance values to be reproducible to within about 5%. Inductor Q for each of the 10 forms is shown in figs. 2 through 11.**

> This data was originally collected by the author in 1969. It has been used since then in many commercial applications, and the values checked with other Q-meters. In all cases the inductance values were within 5% of the predicted value; $Q$ was up to 20% greater, depending on the measuring instrument. **Editor.**

**By Leonard H. Anderson**, 10048 Lanark Street, Sun Valley, California 91352

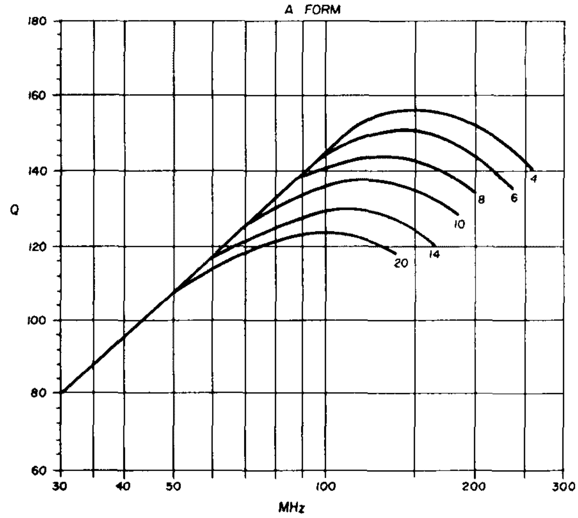


fig. 2. Q of inductors of no. 26 enameled wire wound on 4-40 screw form (curve A in fig. 1).

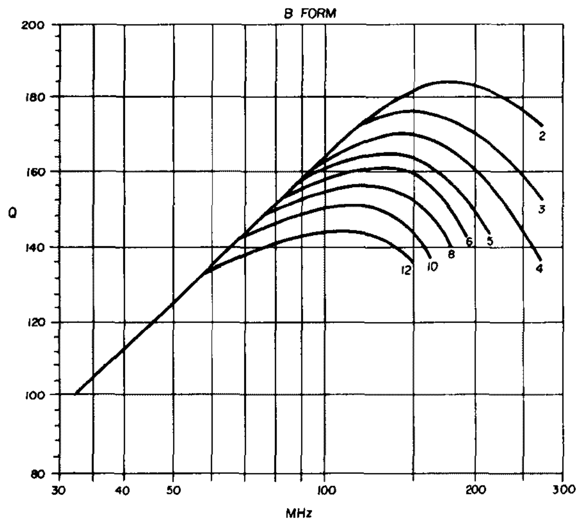


fig. 3. Q of inductors of no. 22 enameled wire wound on 6-32 screw form (curve B in fig. 1).

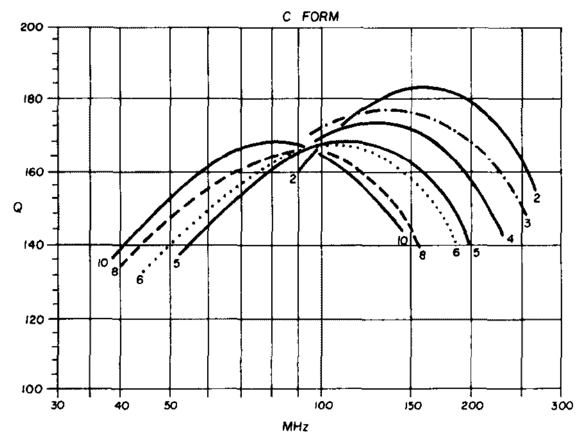


fig. 4. Q of inductors of no. 22 enameled wire wound on 8-32 screw form (curve C in fig. 1).

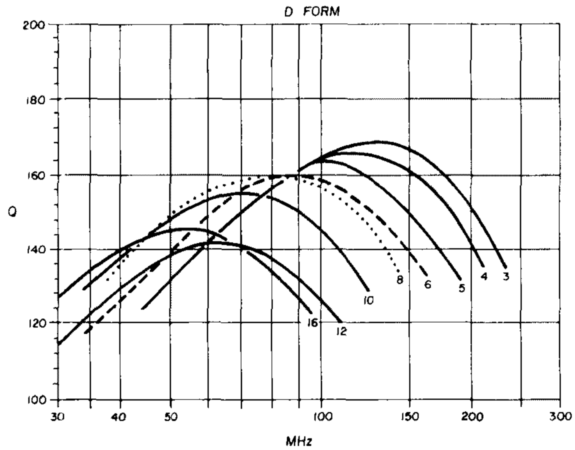


fig. 5. Q of inductors of no. 22 enameled wire wound on 10-32 screw form (curve D in fig. 1).

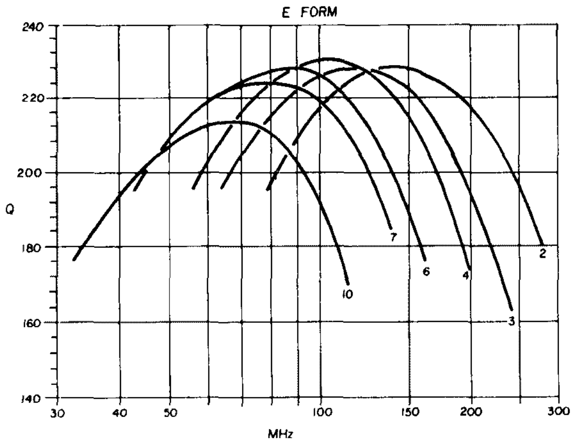


fig. 6. Q of inductors of no. 18 tinned or enameled wire wound on 1/4-20 screw form (curve E in fig. 1).

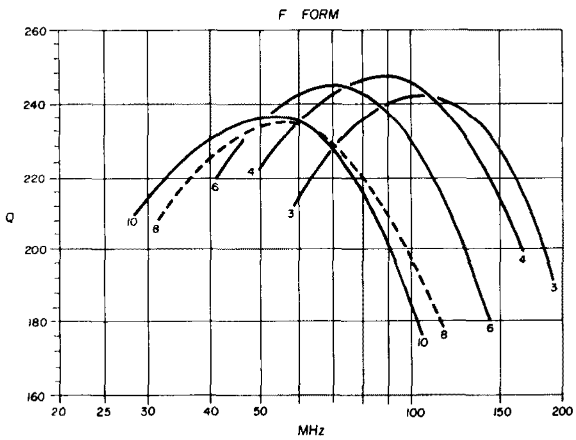


fig. 7. Q of inductors of no. 18 tinned or enameled wire wound on 5/16-18 bolt (curve F in fig. 1).

Wire size is fixed by the number of turns per inch on the screw form. This has been selected to allow a slight space between turns so that adjustment for higher inductance by squeezing may be done if

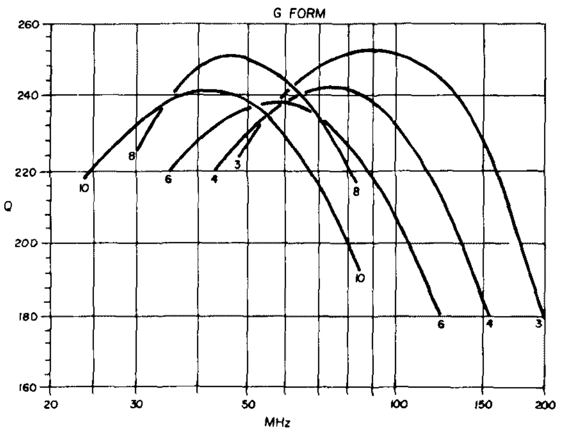


fig. 8. Q of inductors of no. 18 tinned or enameled wire wound on 3/8-16 lag bolt (curve G in fig. 1).

desired. It was also found that wire diameter had to to be restricted to just under full winding by the screw threads.

The no. 12 and 14 AWG wires are household power wires (*Romex*) with the insulation stripped off. If this material is not available, check for wire scraps at industrial plants or at new construction sites; quite a bit is thrown away. The paint-roller ferrule can most often be found on extender poles sold for that purpose and are quite uniform in dimension.

The *Q* curves are shown in figs. 2 to 11. Values in between the indicated number of turns may be interpolated with reasonable accuracy.

All data was obtained by construction and

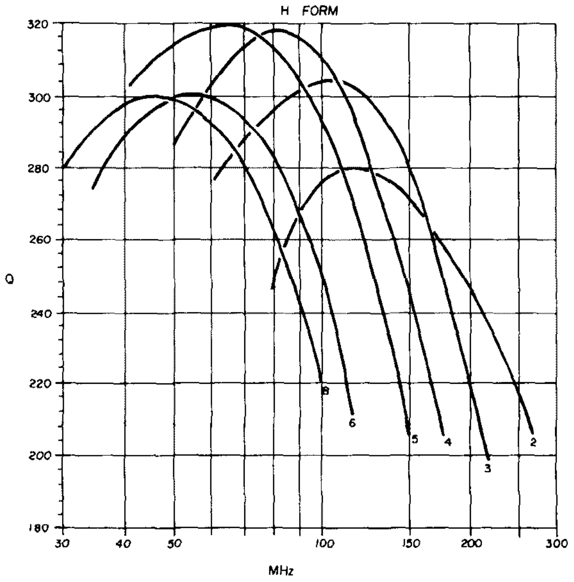


fig. 9. Q of inductors of no. 14 bare copper wire wound on 7-watt Christmas bulb base (curve H in fig. 1).

measurement with a Boonton 190A Q-meter. This limited the measurements to the 20 to 240 MHz instrument range. No compensation for lead length has been made and all coils have an assumed 3/4-inch connection spacing. Each coil was kept at least one diameter away from the top surface of the Q-Meter, consistent with shortest lead length.

Reproducibility of inductance should be within 5% and Q within 20%. This assumes standard coil wire tolerances and lead lengths given. Tolerance may

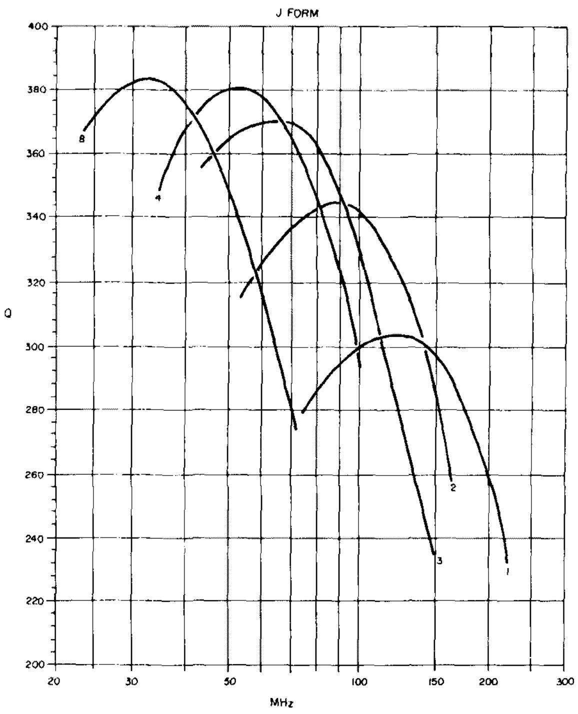


fig. 10. Q of inductors of no. 12 bare copper wire wound on paint roller ferrule (curve J in fig. 1).

drop to 10% for inductors using stripped power wires, depending on the brand; sampling wire from different manufacturers showed 5% tolerances were possible.

Where long leads are necessary, inductance can be adjusted by the following formula for straight wires:

$$L = l[11.7 \log_{10}(0.12 \cdot l \cdot g^2)]\,nH$$

Where: $l$ = *length of wire (inches)*

$g$ = *wire gauge (AWG)*

The formula is an approximation but accurate to 5% for AWG wire sizes from 12 through 26.

All constructions are quite stable and will hold up under most conditions encountered in amateur use. Like all self-supporting inductors made of soft wire,

they are flexible and should not be used in vfo tank circuits or other critical applications.

Bare copper may be coated with varnish to retard oxidation. A light application of spar varnish or polyurethane varnish will lower $Q$ by only 5 to 8 per cent. **Note**: Do note use *Q Dope* since, like all lacquers and acrylics, it will transmit moisture and lift from the non-porous surfaces.

Bare copper holds up surprisingly well. To prove a point, I wound an 8-turn coil on a paint roller ferrule (form J) and buried it in an outdoor planter along with two new plants. The coil was compressed to

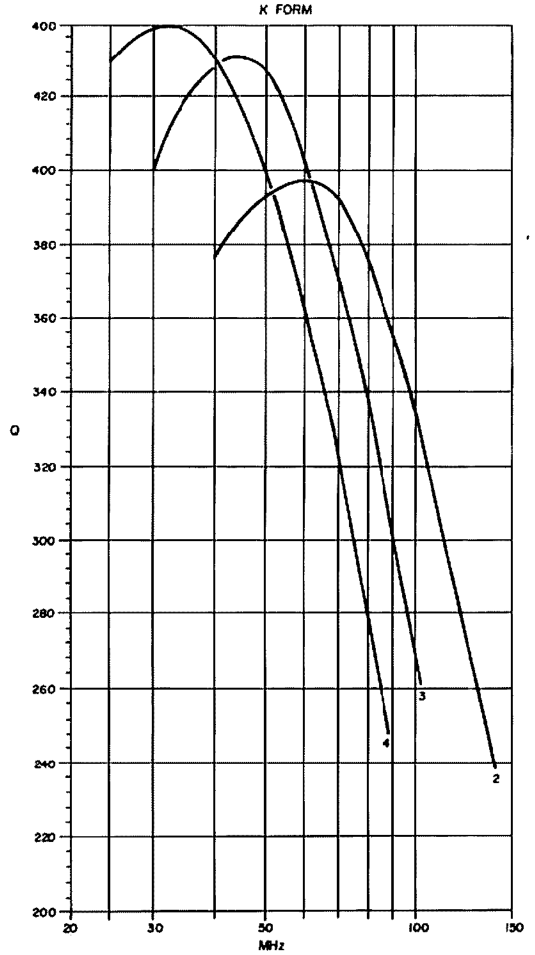


**fig. 11. Q of inductors of no. 12 bare copper wire wound on 117 Vac household lamp base (curve K in fig. 1).**

about 60% of finish length and readings were taken at 50 MHz prior to planting and four months later. Despite watering every other day, the untreated coil had only a 2.3% reduction in inductance and 16% drop in $Q$.

**ham radio**

# gain control IC for audio signal processing

As a multi-purposed IC, the NE570 analog compandor fulfills many audio processing needs

Two **ICs** recently introduced by **Signetics**, the NE570 and NE571, permit the design of efficient and practical audio-signal control functions with a minimum overall parts count. These devices are primarily designed to act as compandors; the complementary processes of compression and expansion.[1,2] They are both dual-channel ICs and either portion can be used individually as a compandor. However, as will be seen in this article they are also well suited to a variety of other tasks useful to the amateur.

## basic device operation

Each channel of the 570 and 571 consists of the functional components shown in **fig. 1A**. Packaged in a 16-pin DIP, the only items common to the two signal channels are the power supply, ground connections, and an internal 1.8 volt bias regulator.

The three principal components of each section are a ΔG cell, full-wave rectifier, and an output amplifier. The ΔG cell is used to control the gain over a range greater than 80 dB. The control voltage for this cell is generated by rectifying an input signal (RECT IN). The final output is then developed by the buffered output amplifier from the scaled signal current supplied by the ΔG cell. The 570 and 571 are identical electrically, but the 570 is selected for lower inherent distortion and a higher supply voltage range.

The ΔG cell, as shown in **fig. 1B**, consists of an op amp, A1, and transistor pairs Q1-Q2 and Q3-Q4. The input signal is first converted by R2 into a current that drives A1. The feedback for this op amp is via the transistor pair Q1-Q2. Therefore, the amount of current in this pair is the same as the current through R2. In addition to driving Q1-Q2, the op amp is also connected to Q3-Q4. Unlike Q1-Q2, this transistor pair does not have a constant-current source. By scaling the Q3-Q4 emitter current, their output is a linear product of the input signal from A1 and the scaled current. This circuit is a linearized transconductance multiplier[4-7] which cancels the inherent non-linearity and temperature sensitivity of the differential pairs, greatly enhancing the usefulness of this gain-control technique.

The rectifier portion consists of op amp, A2, class-B transistors Q5-Q6, a pnp current mirror Q7, and an npn current mirror Q9. When rectifying a signal at the RECT IN terminal, Q5 and Q6 produce pulses of current proportional to the positive and negative input signal swings. The output current of Q6 is used directly, while the Q5 current is mirrored by Q7. Thus, the drive to Q9 is a positive going, full-wave rectified pulsating dc. These pulses are filtered by an external smoothing capacitor attached to the $C_{RECT}$ terminal.

The output stage is a simple inverting op amp similar in performance to a 741. Various options are possible by use of either R3, external input, or feedback resistors. The overall circuit gain is unity (ΔG IN to OUT), with R3 connected as a feedback resistor and 70 μA rectifier current into Q9.

In addition, the THD TRIM terminal allows a small offset to be introduced into the ΔG cell to null its distortion. The two input op amps (A1 and A2) are connected to the internal 1.8 volt regulator. Each op-amp input should be capacitively coupled while the input impedance is determined by R2 or R1, respectively. Circuit operation is very stable and immune to power-supply variations. A single supply voltage from +6 to +18 volts (571) or +6 to +24 volts (570) can be used, though the following applications will use a +15 volt supply.

## basic compandor circuits

The 570 and 571 can be quite simply connected for their basic functions of expansion and compression, as illustrated in fig. 2. These circuits will not be dealt with in great detail because most amateurs will probably be more interested in some of the other uses. Also, compandor operation is covered in detail in other literature.[1-3]

The gain through the expandor shown in **fig. 2A** is $1.43\ V_{IN}$, where $V_{IN}$ is the *average* input voltage.

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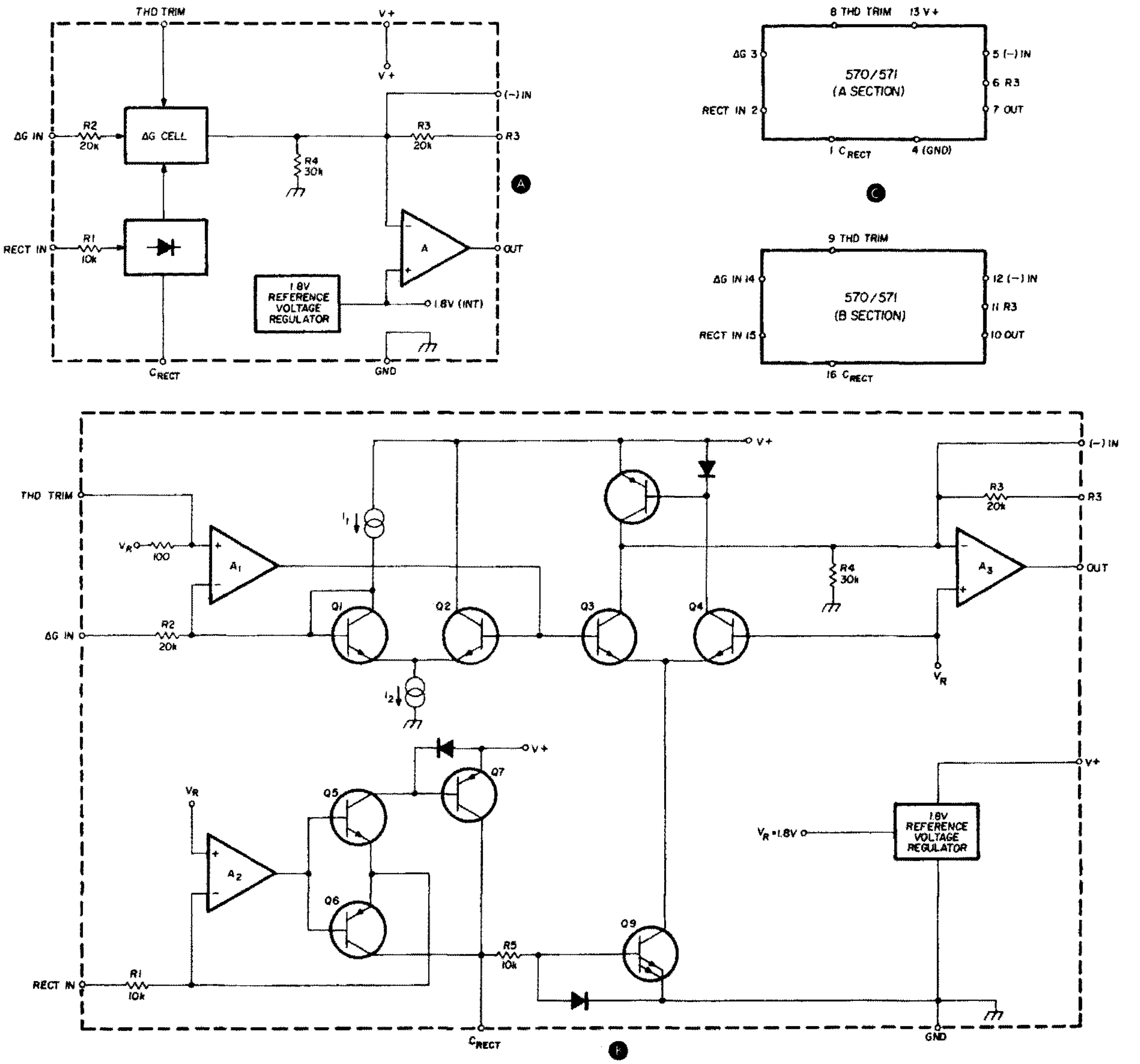



**fig. 1. Functional diagram of the Signetics NE 570/571. A simplified schematic diagram of the device is shown in B. Constant current sources $I_1$ and $I_2$ feed transistor pair Q1-Q2.**

The 570/571 circuit constants are set up such that unity gain occurs at an rms input level of 0.775 volts, or 0 dBm in 600-ohm systems. The $C_{IN}$ and $C_O$ are coupling capacitors, chosen for the desired low-frequency rolloff. $C_{RECT}$ is selected for the desired time constant (10 ms) in conjunction with the internal 10-kilohm resistor (R5).

Resistors $R_A$, $R_B$, $R_C$ and $C_B$ are not essential to basic operation, but are desirable. $R_B$ furnishes short-circuit protection for the output and capacitive load buffering, while $R_A$ and $R_C$ polarize $C_{IN}$ and $C_A$. $C_B$ is a power supply bypass, typically an aluminum electrolytic.

The compressor configuration in **fig. 2B** also has unity gain at 0.775 volt (rms) input, but, a complementary in/out characteristic. The main difference in this circuit is that the ΔG cell is connected as a feedback impedance via $C_F$, and the input is applied to **R3** through $C_{IN}$. Bias for the output stage is set up by the RC-decoupling network, with the values shown appropriate for 15-volt power supply.

In general, the OUT terminal should be biased to one-half the supply voltage. Use of a 570 or 571 as a compandor is not limited to the gains shown, but may be extended to other ranges by use of additional components.

## trimming techniques

Device performance can be enhanced by judicious trimming, as shown in **fig. 3**. Each technique is op-

tional, and can be applied in any combination when the highest performance is desired. The most useful of the three methods is probably the THD trim, which minimizes the gain cell harmonic distortion. In this case, a small voltage (0 to 3 volts) is used to inject a current into the THD TRIM terminal through the 100k resistor. By biasing the rectifier terminal as shown, the inherent current flow in the rectifier is compensated for and permits better low-level signal tracking. Typically, the gain-control signal to A1 should not be reflected in the output. The control feedthrough trimmer will minimize that signal during periods of low input voltage.

## applications

An interesting and versatile group of circuits, the gated or switched-mode amplifier, can be built from the 570/571. With the device controlled by external logic applied to the RECT IN input, the *on* gain is normally set to any value and the *off* attenuation can be in excess of 80 dB. Use of the 570 *or* 571 is advantageous in that all portions of the function can be performed entirely within the IC. Further, the on/off transition times can be set to a value determined by the time constant from $C_{RECT}$.

**Fig. 4A** is a logic controlled amplifier configured for a HIGH input to be on, and LOW off. When the control input is HIGH, CR1 is off and the current

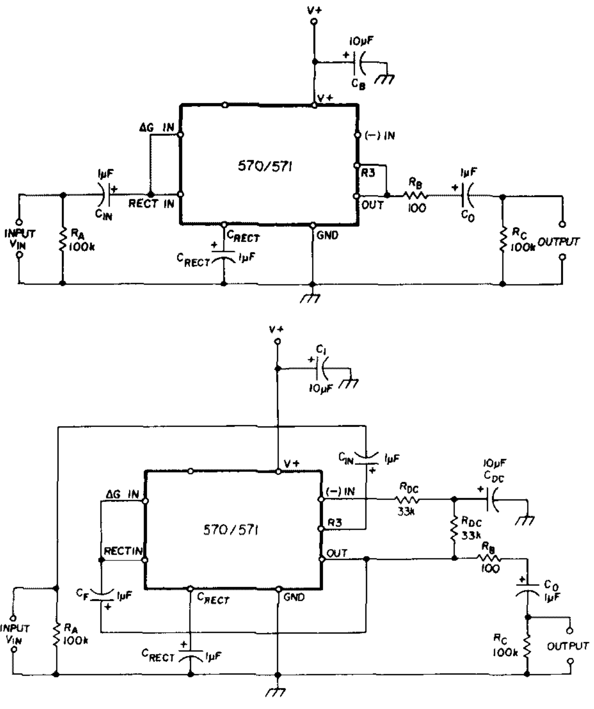



**fig. 2. Schematic diagram of the devices connected as an expander, A, and a compressor, B. The voltage gain through the expander is $1.43 \cdot V_{IN}$, while for the compressor it is $\sqrt{0.7/V_{IN}}$. $V_{IN}$ is the average input voltage.**

developed by $R_{GAIN}$ flows into the rectifier input, which turns on the ΔG cell allowing the signal to be amplified. $R_{GAIN}$ can be selected for the desired *on* state gain, which is unity with a rectifier current of 70 μA. R1 and R3 also effect device gain, but R3 is selected basically for an optimum output bias of 7.5 Vdc. R1 can also be adjusted for gain, but as shown the value allows up to 3 volts rms input/output signal levels.

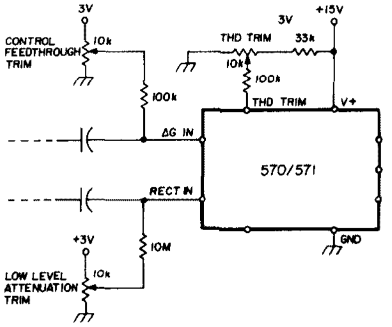


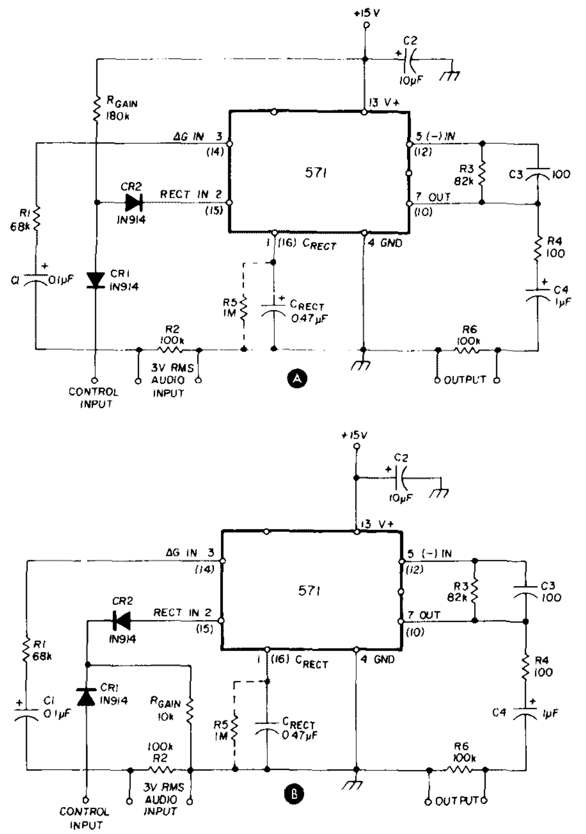

**fig. 3. By applying the different trimming methods, distortion through the 570/571 can be reduced. Though each method is optional, they can be applied in any combination.**

As can be seen in the control characteristics plotted in **fig. 4C**, the gain is unity (or its nominal value, if chosen otherwise) for control inputs greater than 3 volts. Switching is quite abrupt, with full attenuation being achieved at levels less than 1.5 volts. This narrow transition width and the nominal dc center of 1.8 volts allows direct control from CMOS, TTL, DTL, or other positive logic. The ultimate voltage of the HIGH state is non-critical, due to the 100-volt rating of CR1. Unfortunately, this circuit has one inherent weak point. Gain is sensitive to supply voltage due to the connection of $R_{GAIN}$. Thus, the supply voltage should be stable while choosing $R_{GAIN}$ for 70 μA into the RECT IN terminal.

A companion circuit with complementary control characteristics is shown in **fig. 4B**. In this case, the gain is determined by the current developed through $R_{GAIN}$ in conjunction with the internal voltage reference (1.8 V). With a low control input, the normal current will flow out through $R_{GAIN}$. When the control signal is high, CR1 is forward biased, interrupting the current flow. Therefore, the output will be attenuated since Q3-Q4 have been turned off.

Both circuits can be tailored for specific on-off transition times by selection of $C_{RECT}$. The time constant is simply $10k \cdot C_{RECT}$ (10 kilohms is the internal resistor). Thus, the audible switching effect can be smoothed, eliminating the transients produced by an asynchronous fast switch. The $C_{RECT}$ value shown yields nominal times of 5 milliseconds.

Use of $C_{RECT}$ in a switched amplifier of this type is



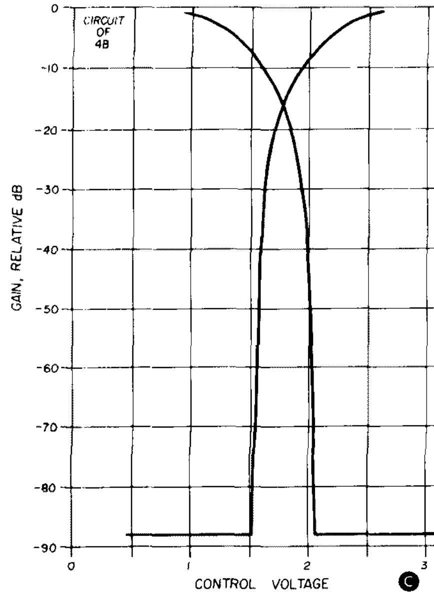




**fig. 4. The logic controlled amplifiers can be configured to provide an output with either a HIGH or LOW input. This allows the amplifiers to be interfaced with many different logic types, TTL, CMOS, DTL, etc. The resistor $R_{GAIN}$ should be selected to provide 70 $\mu$A at unity gain. Each amplifier has an on/off time determined by the time constant of $C_{RECT}$ with internal 10k resistor.**

optional and not absolutely necessary. However, to minimize noise pickup some capacitance will be found useful. Also, the ultimate *off* state attenuation will be limited to about 60 dB due to the internal bias current. This *feedthrough* error can be eliminated by connecting a 1 megohm resistor from $C_{RECT}$ to ground to bleed away the error current. This allows attenuation of 80 dB or more.

**Fig. 5** illustrates two sections of a 571 combined as a two-input multiplexer, for FSK or other uses. This circuit operation is similar to the others, but is biased and switched in a simpler manner. Gain of each *on* channel is unity, as determined by $R_{GAIN}$. The output of the B channel ΔG cell is summed with channel A by connecting the (−) IN terminals of the A and B sections. The respective channels are gated *off* by a low control logic input, which clamps the rectifier current, switching the ΔG cell off. For fsk or alternate channel use, the CONTROL A and CONTROL B signals should be complementary. Thus, the input is "instantaneously" switched between the **A** and **B** inputs.

Control signal suppression can be optimized with the CHOPPER NULL control, which trims the control signal component in the output. Suppression is better than 60 dB after trimming. Response time is quite fast, and is actually limited by the slew rate of the output op amp rather than the ΔG cell itself. This makes the switching interval a function of the signal's peak amplitude. For instance with a 4-volt peak amplitude signal the 0.5 V/$\mu$s slew rate will

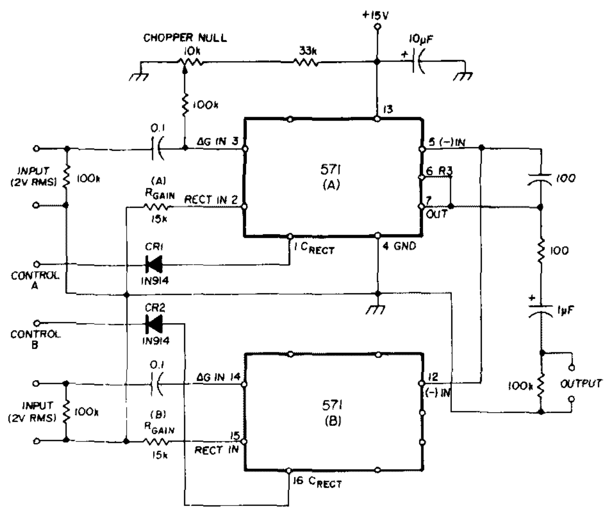



**fig. 5. By providing complementary control signals, the FSK generator will switch between the two signal inputs. The outputs, when ON, are summed through the first operational amplifier.**

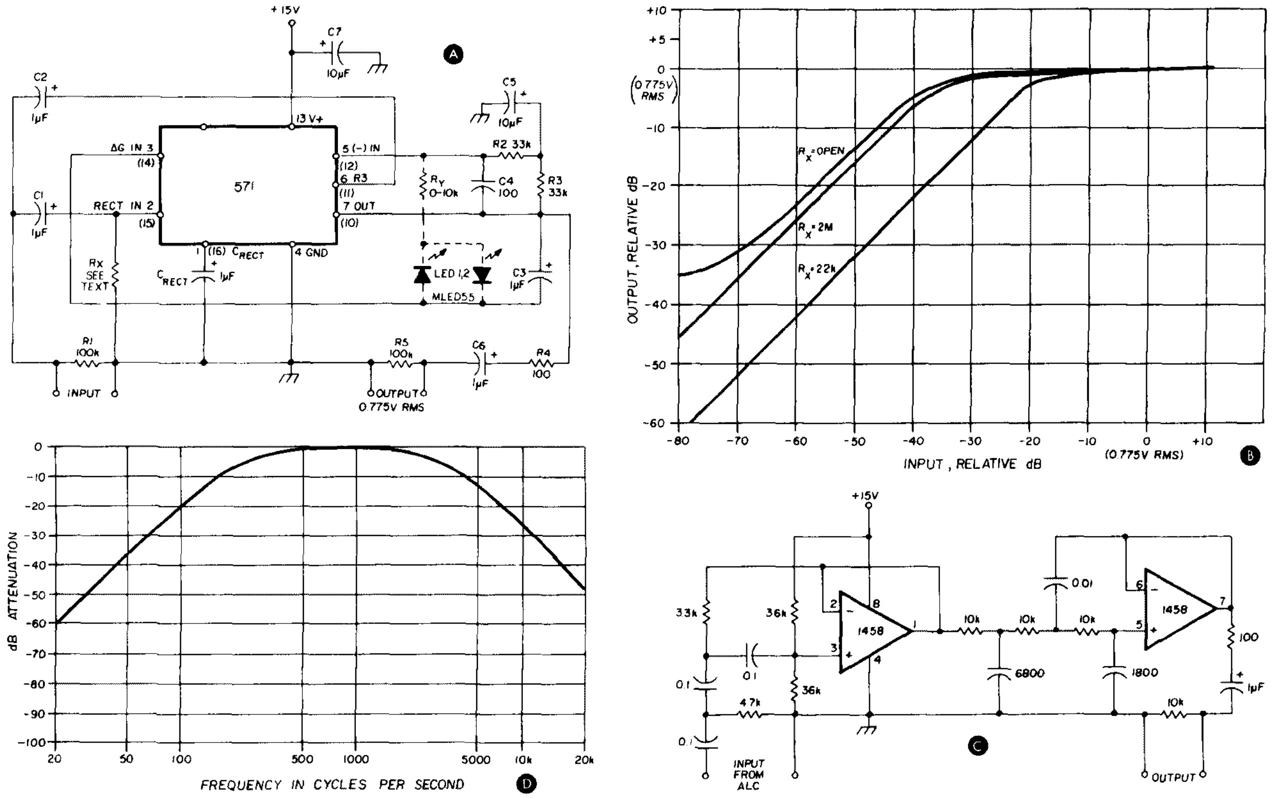



**fig. 6. Schematic diagram of an NE670 connected as an automatic level control (A). The speech filter (C) is added to remove the undesired components created by the peak clipping. The attenuation curve (D), though not sharp, provides excellent rejection of the sharp pulse spikes generated by the clipping.**

allow switching in 8 μs; lower amplitudes will be proportionally faster. Although the circuit is touted as a multiplexer, it can also be used as a summing switch, with both signals on at any given instant.

## automatic level control

Automatic level control is a relatively common requirement in audio signal processing. A 571 automatic level control circuit can provide constant, high percentage modulation with varying input levels, yet without danger of overload if properly handled.[8] This circuit (**fig. 6**) is adapted from the 570/571 data sheet. There is one additional feature which may be useful, however: an optional resistor allows the threshold of level regulation to be varied. If $R_X$ is left open, the circuit will have its widest range of gain control. As this resistor value is lowered, a larger input signal is required for full output. The general effect, for various $R_X$ values, is shown in **fig. 6B**.

Since the 570/571 device operates on the principles of *average* level detection, it will easily saturate on large crest factor input levels such as speech, or large transient envelope changes. This condition is highly undesirable, but, fortunately its effect can be negated quite simply. Since the result of overshoots are peak-to-peak amplitudes in excess of the regulated output level, it follows that appropriate peak-level clipping can effectively control the overshoots. In this case, the rms output amplitude is 0.775 volt or 2.2 volts p-p. This particular level is conveniently clipped with a pair of reverse paralleled LEDs, which will limit to 3.2 volts p-p. The LEDs can be connected as shown in **fig. 6** with a series resistor $R_Y$, which is used to regulate the clipped amplitude.

This clipping technique, while not a requisite part of the automatic level control, greatly enhances its regulation with nonsinusoidal signals. Use of this circuit with speech inputs will necessitate the diodes, which will typically be clipping a good portion of the time. This, of course, adds audible distortion to the output. Therefore, a useful item with the ALC/clipper is a speech filter to remove the superfluous high and low frequencies. The filter will also greatly attenuate harmonic components generated by clipping.

The circuit in **fig. 6C** is a bandpass speech filter which uses a single IC. The circuit is simply a pair of cascaded Sallen and Key[10] highpass filters (3-pole Bessel type). The Bessel response is one of the poorest in terms of cutoff sharpness, but good from

a pulse response standpoint. This feature is important to minimize amplitude overshoots which could occur with severely clipped inputs.

A common 1458 (dual 741) op amp is used with nearest 5 per cent component values for the filter elements. If low-power operation is desired, the 1458 can be replaced directly with a 358. If a 358 is used, 10 kilohm resistors should be added from each output terminal to common. With unity gain, the circuit can drive load impedances greater than 10 kilohms.

One very effective use for the 570 and 571 device is an amplitude-regulated RC sine wave oscillator. Typically, such circuits use a Wien bridge or other frequency-selective RC network, with some form of amplitude stabilization to maintain constant and correct loop gain, and also to guarantee output waveform purity. A 570 or 571 is nearly optimum for this type of circuit because it contains the required functions of amplifier, rectifier, and gain-control circuits.

Two types of sine wave oscillators are shown in **fig. 7**. The oscillator circuit (**fig. 7A**) based on the Wien network is formed by the combination of R1-C1 and R2-C2. This network is placed around the output amplifier of section A, which effectively makes it a bandpass amplifier resonant at

$$f = \frac{1}{2\pi RC}$$

With equal values of R and C, the input/output voltage ratio is 2 to 1.

To originate and sustain oscillations, the 571B section is used as an inverting amplifier with a nominal gain of 2. A slightly greater initial gain is established by the combination of R6, R7, and R8, which ensures startup. The B section ΔG cell is connected as a compressor, which regulates this stage's gain at the precise value required to maintain undistorted, stable amplitude oscillations.

There are two main steps taken to enhance flexibility of the circuit. A separate dc feedback path (R3, R4, C4) is used around the A stage, to remove value restrictions on R2 due to bias considerations. This allows R2 (R1) to range from 10k to 1 megohm without a major performance compromise. C1 and C2 have an even greater range, from 1 μF down to 100 pF. To minimize error due to strays, the lowest value should be used. With the values shown, the circuit is capable of reasonably low harmonic distortion. For example, 0.03 per cent distortion was measured at 1.6 kHz and THD (Total Harmonic Distortion) can generally be held below 0.1 per cent. This will vary according to the specific frequency, and the selected impedance of the Wien network. The low value of distortion is due to the light degree of ΔG cell regulation.

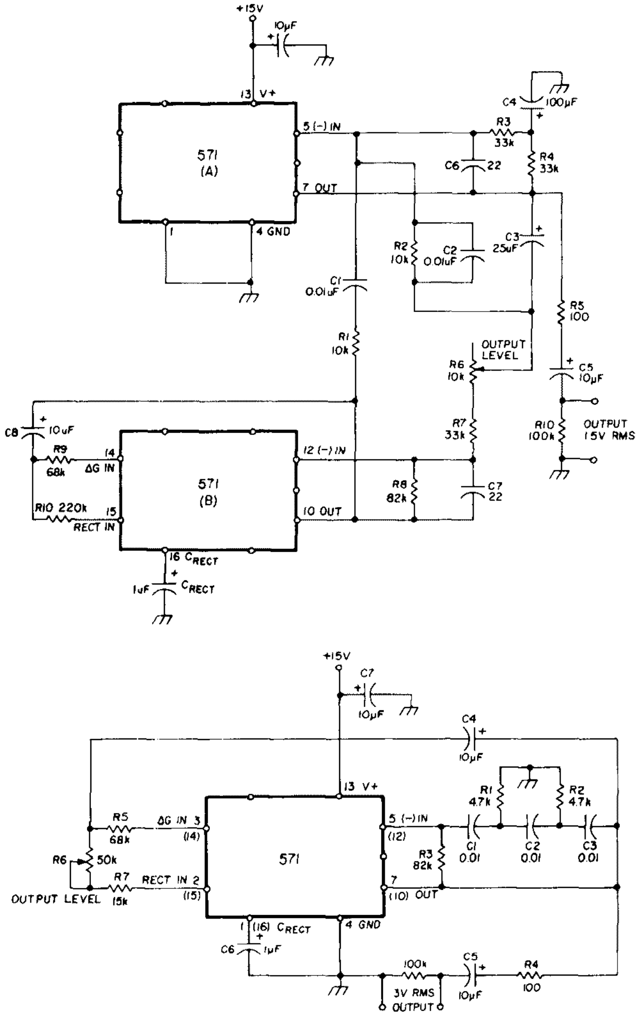



**fig. 7. The NE570/571 can be connected as a sine-wave oscillator. The Wien bridge type oscillator is shown in A. For R = R1 = R2 and C = C1 = C2, the operating frequency is 1/2πRC. Resistor R should be limited between 10k and 1 megohm with C between 1000 pF and 1μF. The normal frequency range can be varied from 10 Hz to 10 kHz. The phase-shift oscillator should be used to generate discrete frequencies only. Depending upon the selection of parts, the output frequency will be 1/2πRC√3.**

The circuit will operate as shown over the range from 10 Hz to 10 kHz. Below 10 Hz component size becomes impractical, and above 10 kHz slew limiting in the output amplifier causes distortion to rise. The circuit is useful as a fixed frequency oscillator, but can also be tuned if a matched dual pot is available for R1-R2. Output amplitude is set by R6, and is optimum at 1.5 volts rms output, from section A. If a higher output level is needed, section B output can also be used, at 3 volts.

The circuit of **fig. 7A** may be unduly complex for some uses, so an alternate and much simpler sinusoidal oscillator is shown in **fig. 7B**. This circuit is a form of phase-shift oscillator, similar to that

described by Tobey, Graeme, and Huelsman.[9] A 571 is well suited for a phase-shift oscillator because it contains the necessary inverting amplifier to sustain oscilla tion. In the circuit shown, C1, C2, and C3 are the timing capacitors, while R1 and R2 are the resistors for the phase-shift network. R3 must be at least 12 times the R1-R2 value for adequate loop gain. AGC is provided by using the $\Delta G$ cell as a compressor.

This circuit is not suitable for tunable use. It should only be used as a spot frequency oscillator, by varying C1, C2 and C3. This is because R1 and R2 are related, by the design, to R3; in this specific case R3 cannot be variable because it is used to set the output dc bias point.

Although it uses a simple design, this circuit produces excellent results. At the frequency indicated, a laboratory test indicated a THD of 0.01 per cent at 3 volts output, which is remarkable in view of the circuit's simplicity. To take full advantage of this performance, an output buffer may be useful; for this you could simply use the remaining channel as a simple unity gain inverter.

## conclusions

This discusssion has covered a few uses for a new and interesting chip. In the course of this article's preparation several other potential uses suggested themselves, such as phase comparators, phase-locked loops, voltage-tuned oscillators, and others. Unfortunately, space and time restrictions did not permit their complete examination.

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**ham radio**

# high dynamic range two-meter converter

Circuit details for a 2-meter converter with a +15 dBM intercept point and 5 dB noise figure

**This two-meter converter** is an improved version of an earlier model[1] and is a result of research and development which I did around 1969 at AEG-Telefunken in Ulm, Germany. The German Ministry of Postal Affairs (equivalent to the U. S. FCC) requires that for all approved receivers for commercial and military application, a dynamic adjacent channel measurement must be performed. The result of the test depends upon the third-order intermodulation-distortion characteristic of the input stages.

A Radio Amateur group has used this particular converter circuitry which was later referred to as the "Martin front end."[2] This technique is also successfully used in the Rohde & Schwarz high-frequency Communication Receiver EK47, and was later adopted by Southcom and Atlas. The basic purpose of this circuit is to get the best performance out of a mixer by obtaining the lowest possible noise figure. To achieve this, the i-f output circuit has to be properly terminated over a large frequency range. This feature was previously described in *ham radio*.[2] During the last few months, the third-order intermodulation distortion has been evaluated in several magazines. However, a so-called *second-order* intermodulation distortion problem has received little mention.

A two-tone test is used for both second- and third-order IMD performance measurements. Second-order performance is checked at $f_1 \pm f_2$. Third-order is the performance at $f_1 \pm 2f_2$.[5] To reduce the effects of second-order IMD, it is necessary to use as much selectivity as possible and then compensate for the losses of these filters by using appropriate amplifiers. A suitable low-noise preamplifier with wide-band matching will be discussed later.

The *two-meter* converter shown in **fig. 1** combines these techniques. It was designed for extreme linearity and selectivity with the added goal of keeping the noise figure below 5 dB.

Many converters with noise figures below 3 dB described in the literature use neutralization. This method has two distinct disadvantages in the way it is currently done:

**1.** The neutralization can be made only over an extremely narrow frequency range, and

**2.** Since, in most cases, amateurs do not have ade-

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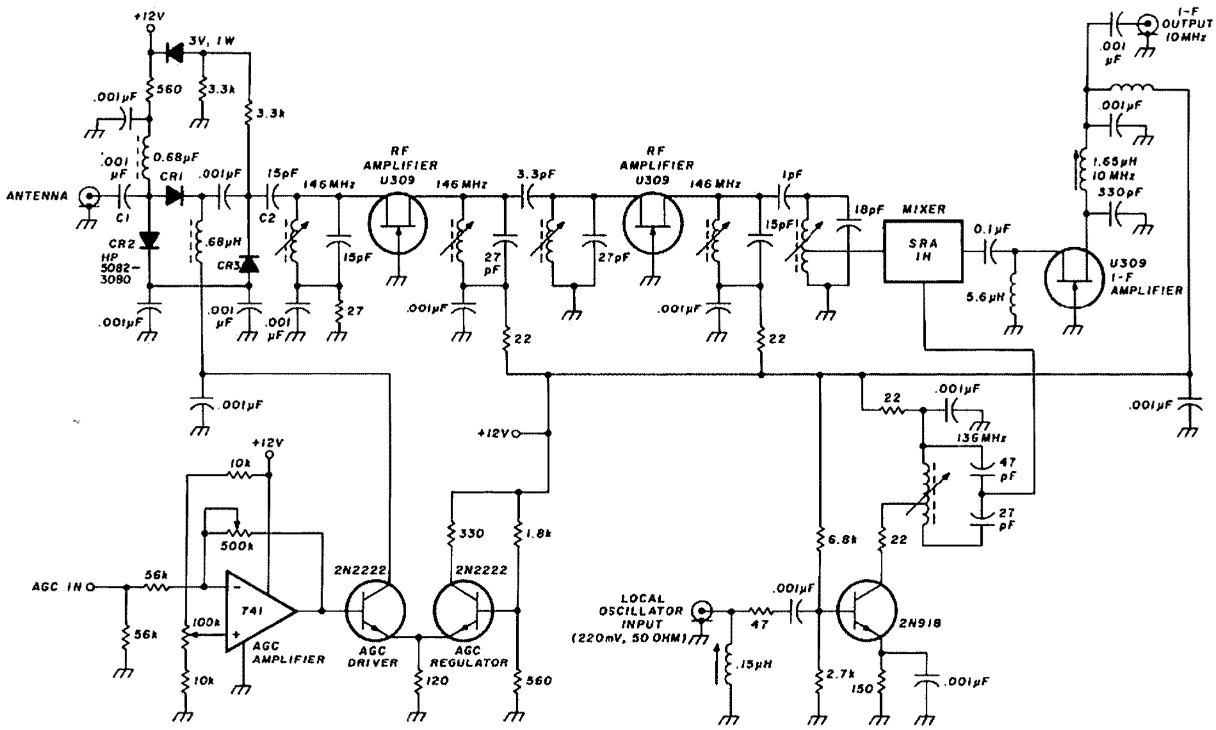
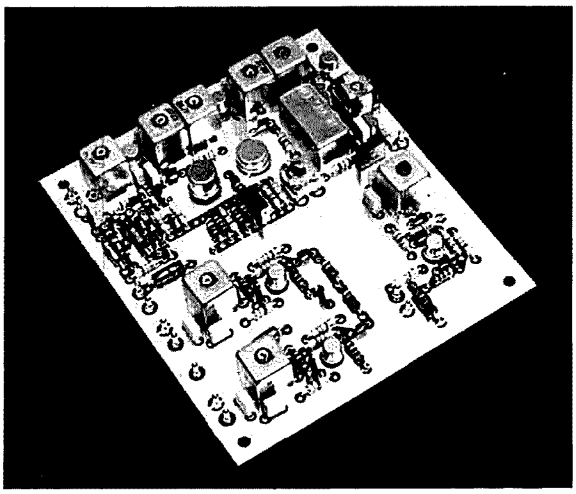


**fig. 1. A two-meter converter with +15 dBm intercept point, 16 dB power gain, and less than 5 dB noise figure.**

quate test instruments, an exact adjustment is seldom achieved, which degrades the two-tone IMD performance.

This converter has five tuned circuits at the input frequency which results in an overall bandwidth of barely 4 MHz. Proper tuning of this converter can best be achieved by the use of a suitable sweep generator.

These five tuned circuits provide an image suppression of 60 dB for an i-f of 10 MHz, or more than 80 dB for an i-f around 30 MHz. To simplify the circuit, U-309 fets are used which should have an $I_{DSS} = 20\ mA$. The third-order IMD of these transistors can be neglected as compared to the performance of the SRA-1H mixer. The overall gain between the antenna termination and the mixer input is about 10 dB. Therefore, the overall intercept point of the converter is +15 dBm with a noise figure of slightly less than 5 dB. Part of the fairly high noise figure is due to the 1-dB loss in the pin-diode attenuator.


**Photograph of the converter board. Two additional output stages have been added.**

This converter uses the ground-gate field-effect transistor circuit as described in reference 2. However, in the original version,[1] a special bipolar transistor in grounded-base configuration was used and provided a suitable wideband match for the mixer. This converter is shown in fig. 2 as part of a transceiver that uses a frequency-locking system to stabilize the free-running oscillator in increments of 6.25 Hz (semi-synthesized).

## unconditionally stable low-noise input stage

The low-noise preamplifier shown in fig. 3 is based upon a circuit suggested by AEG-Telefunken in the 1950s and first published by me in English.[3,4] It was used to achieve extremely low-noise input stages with triodes while avoiding neutralizing circuits with

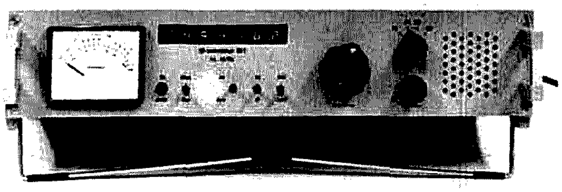


fig. 2. Photograph of a semi-synthesized 2-meter transceiver using the converter shown in fig. 1.

their inherent mass production problems. The same feedback arrangement not only avoids all instability problems but also improves the dynamic range. With practically all other neutralizing circuits a certain degree of performance reduction is observed.

The amplifier shown in **fig. 3** is a mixture between a grounded-gate and a grounded-source circuit. It is a bridge arrangement which neutralizes the feedback

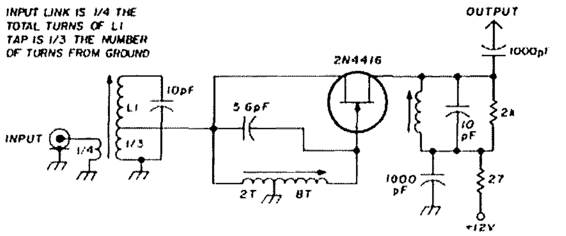



Fig. 3. Low-noise pre-emplifier with rf feedback to provide unconditional stability and low distortion.

capacitance between gate and drain. In addition, the input impedance (*I/S*) is transformed in parallel between the gate and ground and provides the necessary wideband characteristic. A noise figure of between 1 and 2 dB, using the inexpensive 2N4416, can be easily achieved and the gain is roughly 15 dB. The circuit is unconditionally stable. This is the only circuit known to me which combines optimum matching for best noise, lowest input swr, and best matching for high power gain.

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**ham radio**

# caption device for slow-scan television

## An easy way to dress up your video with professional-appearing captions

**I built** a camera for slow-scan television (standard 625 line) and needed a method to rapidly change the material being televised. Because of space limitations, the camera had to be located fairly close to the subject. Using an $f1.9$, 16-mm focal-length lens resulted in a subject area of 4 x 4 inches (102 x 102mm) with the camera 9 inches (230mm) away, giving a 3¼-inch (83mm) square reproduction on my monitor, which uses a 5-inch (127.5mm) type 5FP7 tube.

On the receiving end the monitor-screen size is unimportant since, with the 128-line system used, the screen will be filled depending on amplitude-control settings.

### description

The caption device is based on the use of plastic-strip magnets. The type I used are strip magnets for refrigerator doors; however, small bar magnets about 0.2-inch square by 1-inch long (5mm by 25.5mm) work very well.

A sheet of 16-gauge steel plate 9-inches (230mm) square was used as a backing plate to hold the caption board (described later). The backing plate was mounted on a frame so that it was square with the camera. The backing plate can be painted a matte black or covered with plastic self-adhesive material, which is obtainable at most hardware stores.

The next step is to make the caption boards. Two boards were made, one for letters and numbers and one for prepared pictures and captions. These are called the "variable caption board" and the "prepared caption board." Each is described separately.

### variable caption board

For this board I obtained a piece of 0.1-inch-thick (2.6mm) card stock and cut it to 9.5 by 5.5 inches (242 by 140mm). The size depends on the lens you are using, but this size was for the $f1.9$, 16-mm lens mentioned previously.

I then cut the plastic magnet material into strips about 0.15 to 0.2 inch (3.8 to 5mm) long using a fretsaw. If you can obtain magnets near this size, so much the better. The objective is to obtain strips of magnetic material measuring 0.15 to 0.2-inch (3.8 to 5mm) square.

The card was marked as shown in **fig. 1.** Slots were cut using a sharp modeling knife and straightedge. The card was then put onto a flat surface and the magnets inserted into the slots so that the two original surfaces of the magnets were face down and face up, respectively. The rough-cut edges should mate with the walls of the cutout slots.

The magnets should now project from the card on the side facing you but will be flush with the face of the card on the other side. While the card is in this position, mix some two-part quick-set epoxy and place blobs around the magnets. When the epoxy is cured, the card can be covered on the flush side with a piece of plain white paper using impact adhesive.

We now have a basic board, plain white on one side with slightly projecting magnets on the other. The board will remain in position when placed against the steel supporting plate.

**Alphabets.** The next step is to prepare a couple of

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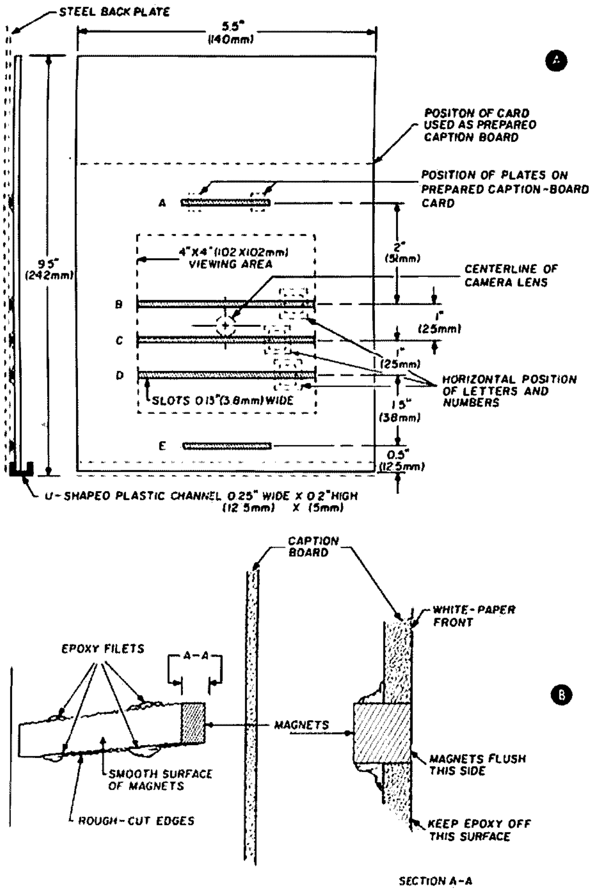



**fig. 1. Caption-board construction, A, and details for mounting the magnets onto the board, B. Dimensions are for the variable caption board. For the prepared caption board only cutouts and magnets marked A and E are used.**

alphabets and numbers to use with the variable caption board. Purchase a sheet of rub-on letters (available from stationery stores or commercial-artist suppliers). Choose a bold condensed type face about 3/4-inch (19mm) high.

Using a piece of clean white card stock, rub on two of each letter in the alphabet and all the numbers. Keep them within the lines as printed on the sheets and try to space them equally for a professional appearance. Burnish letters and numbers using the rub on sheet backing.

Next, carefully cut the characters into identical-height strips (except the letter Q, which must be slightly larger than, say, A or B). Use a square and cut each character from the strip into a rectangle as shown in fig. 2.

Obtain some thin steel washers about 3/8-inch (9.5mm) in diameter. Mark the exact center of the back of each letter card, then affix the washer in position using double-sided adhesive tape. You now will have alphabets and numerals that can be put onto the prepared magnet card in all sorts of combinations. If you cut the letter or number cards so they are square, they should mate evenly.

**Applications.** The magnet board can be made off-camera then placed against the steel supporting plate for on-the-air display. The result will be a neat and flexible method for originating captions. If you can't find the recommended steel washers, small pieces cut from tin-can tops can be used.

## prepared caption board

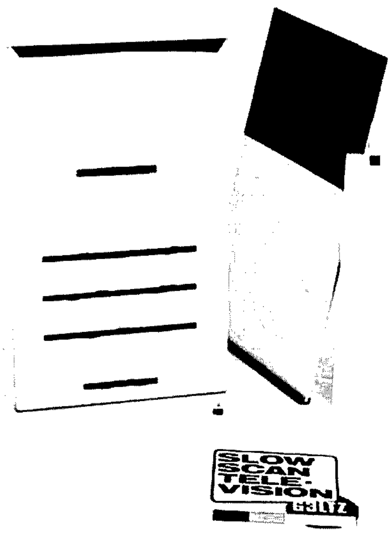
This board is similar to the variable letter and number board but uses fewer magnets. It is prepared in the same way as the variable caption board (see details A and E, fig. 1). The two magnets shown hold the board against the steel backing plate and also hold the prepared material in place.

For this board, cut thin pieces of card stock to 5½ by 7 inches (140 by 180mm). Use a prepared stencil with a 4 by 4-inch (102 by 102mm) cutout. Draw lightly in pencil a 4 by 4-inch (102 by 102mm) outline on each sheet so that it's centered and in exactly the same position each time. (Retain the stencil for future use.)

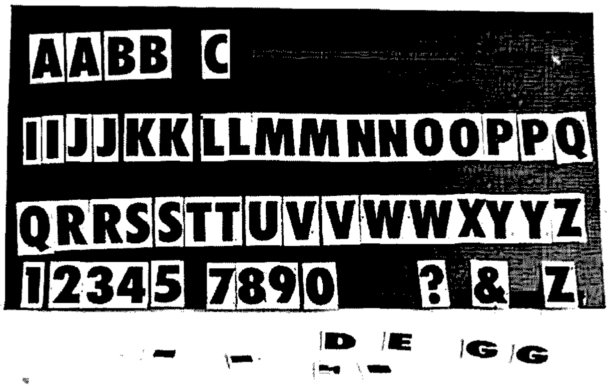
Next, paste up your photos, rub-on letters, or test cards on each sheet within the 4 by 4-inch (102 by 102mm) outline. These sheets will usually contain standard material used on the air such as, "CQ de XXX," "My name is Joe," and so forth.

Returning now to the cardboard with the two magnets inserted, epoxied and covered with white paper, the next step is to make a small ledge along the bottom to hold the bottom of the prepared material. The top is retained by steel washers or tinplate affixed to the back of each sheet next to the magnet inserted in the cardboard. The bottom holding rail is made from a piece of plastic sliding door runner such as used in cupboards. It is held in place with impact adhesive.

The finished board can now be placed onto the steel




**Rear of the variable caption board, left, and front of the prepared caption board, center. The rear of a prepared caption card is shown at right. The two steel wahsers hold the card to tha magnet board. A prepared caption card is shown at the bottom.**




**Storage rack for the caption letters and numbers. It is made of cardboard with strips of magnet cemented to the surface for easy removal of letters. Note the pieces of tinplate on the backs of removed letters.**

backing plate, which holds the board vertical and square. Prepared sheets can be quickly dropped into the bottom rail and pushed gently back. They will remain in position easily.

You can also use this system to televise unprepared material simply by placing your drawing or picture onto the steel back plate. Use one or two bar magnets to hold the material to the plate.

## final remarks

The system has proved its worth in use. With the two-magnet boards and steel back plate, material can be changed quickly and positioned in the same place during on-the-air contacts. Obviously the idea can be adapted to fit your own camera. The dimensions given are such that, with my camera on a flat table, the center of the magnet board 4 by 4-inch (102 by 102mm) viewing area lies in line with the center of the camera lens. A couple of stick-on dots indicate the horizontal position of the magnet board when placing it onto the steel back plate, and two pieces of tape are used to position the camera.

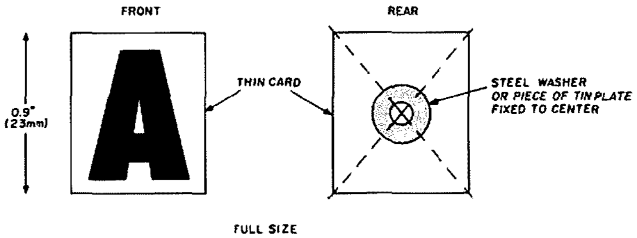



**fig. 2. Construction of individual letters or numbers. The author used Letraset sheets (2683 in Europe; 113-72-CN in the USA).**


## acknowledgement

My thanks to Bob Weston for the photographs.


**ham radio**

# continuity bleeper

## for circuit tracing

The four comparator amplifiers in the Motorola MC3302P IC are put to work in this circuit tester

When tracing circuits with an ordinary continuity tester, the procedure is slowed by the need to look away from the circuit to read the meter. The procedure is also risky because of the inability to distinguish between conductors and low-resistance components, such as inductors. Electromechanical buzzers are better, but they pass currents or induce voltages that damage transistor circuits.

These thoughts led to the design of the continuity tester described here. It delivers a continuous audio tone when the test terminals are connected by a resistance of less than about 1 ohm. The circuit under test never sees more than 3 volts or 300 mA, depending on its resistance. The tester ticks gently when switched on but open-circuited, as a reminder that the battery is on load, although the battery load is only 0.8 mA.

### description

This complex pattern of operation is made easy by the Motorola low-power MC3302P IC, which contains four identical comparator amplifiers (fig. 1). These amplifiers are used in the tester as a measurement comparator, audio oscillator, loudspeaker amplifier, and economy ticker. The unusual differential input arrangement, fig. 1, allows the chip to operate down to zero common-mode voltage with a single supply of only 3 volts nominal. The current-sinking output transistor is particularly appropriate for the measurement comparator, U1A (fig. 2).

### packaging

The MC3302P IC, together with battery and a 2-inch (51mm) diameter speaker, is enclosed within an aluminum box measuring 4¾ x 3¾ x 2 inches (121x96x51mm). The method of construction is shown in the photos.

### circuit

Referring to fig. 2, U1A is a balance detector in a

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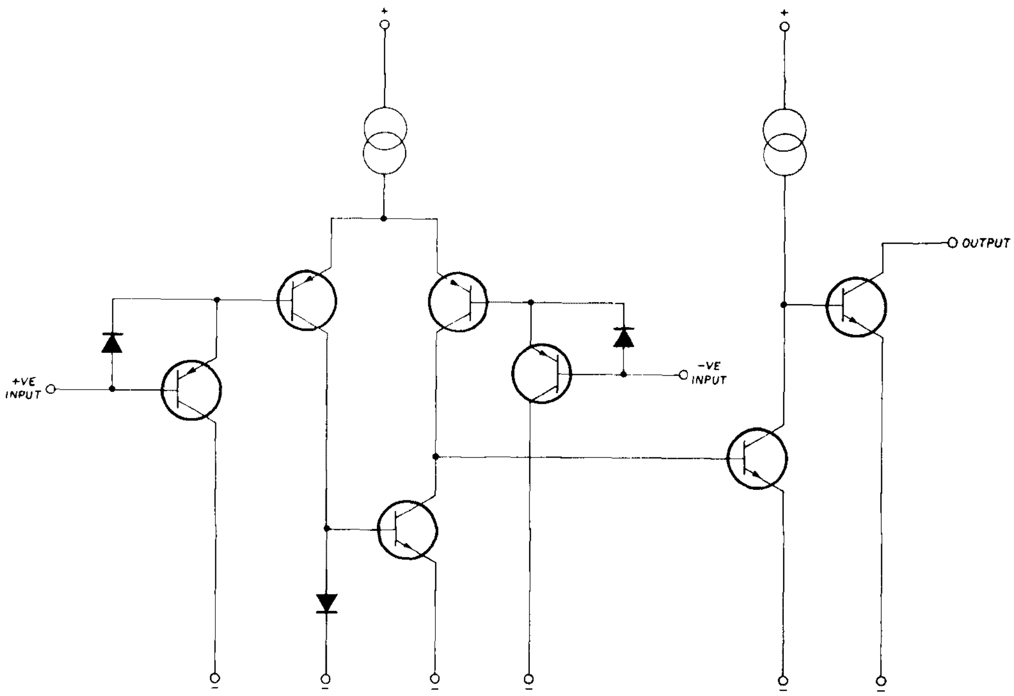



fig. 1. Schematic of one of four identical comparator amplifiers contained in the Motorola MC3302P IC, which is the heart of the continuity bleeper circuit.

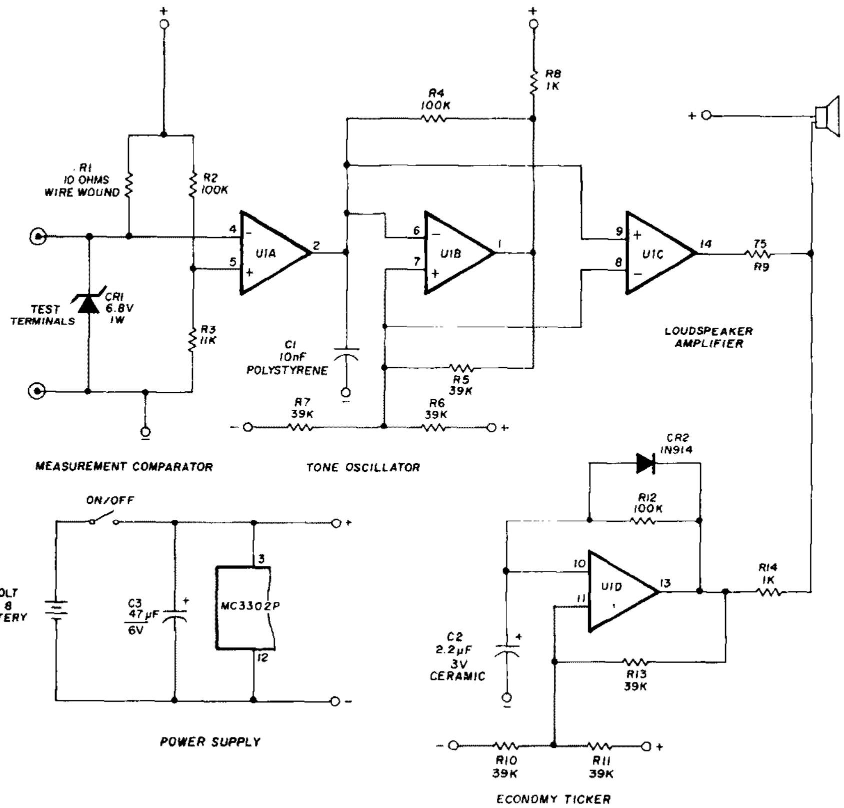
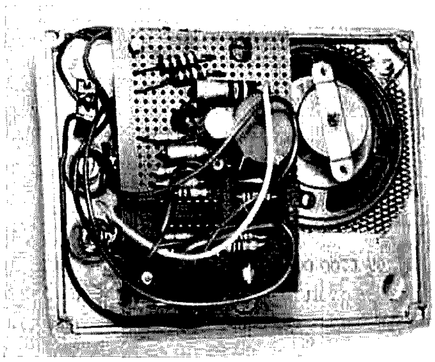



fig. 2. Continuity-bleeper schematic. U1A through U1D, all contained in the MC3302P, provide the four circuits for the tester. Pins 3 and 12 of the IC are used in the power supply. All resistors are ±10% 1/8 watt, except R1.

bridge circuit. If the external resistance between the test terminals exceeds the product $R1 \times R3/R2$, the output short circuits C1 and disables tone oscillator U1B.

Measuring-circuit tolerance is determined by R1, R2, R3 *in the measurement comparator*, and by the offset of U1A. The threshold is between 0.7-1.6 ohms. If desired, R2 or R3 may be trimmed to set the threshold to, say, exactly 1 ohm or 0.5 ohm. Power zener CR1 protects the tester should it be connected to a "live" circuit.



**Entire circuit including speaker and power supply is enclosed in a small aluminum box.**

Comparator U1B is a 600-Hz oscillator. Resistors R5, R6, R7 provide positive feedback, so that comparator output snaps rapidly through a range nearly equal to the supply voltage. This output is returned to the other input through R4 and C1, whose voltage is an approximate sawtooth that causes the comparator to switch when the voltage increases to 2/3 and falls to 1/3 of the supply voltage, thus generating the square-wave output.

The input of speaker amplifier U1C is connected to ensure miminum supply current when U1A turns off the tone oscillator. Resistor R9 defines volume and current consumption when the tone is on.

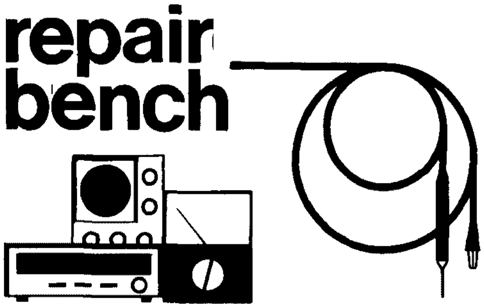
The economy ticker, U1D, draws a 0.25 millisecond pulse of current through the speaker every half second. Its operation is similar to that of U1B, apart from the extra diode, CR2, which gives the short discharge time for C2 that corresponds to the "tick." If the tick isn't loud enough, resistor R14 may be reduced to 220 ohms (at the cost of increasing current consumption to 1.8 mA).

### power supply

The battery must be able to deliver 0.3 ampere to the measuring circuit, so an inexpensive two-cell lantern battery is used in this application. Battery replacement will be rare, so the battery is secured between the supports of a Veroboard component panel. Connection is by insulated spring clips that bridge these supports.

With the availability of the MC3302P, the circuit described here has become a practicality. It can be a great time saver for both experimenter and professional in circuit testing.

**ham radio**




# Joe Carr, K4IPV

## simple antenna instrumentation

**Regardless of the type of antenna** you erect, it will rarely be absolutely correct despite the fact that instructions were followed carefully and textbook formulas were used. This is because such formulas are either based on an ideal situation or use certain assumptions that may or may not be valid in your particular case.

Fine tuning an antenna system is all but impossible without some sort of instrumentation to tell you what's going on. In fact, it's wise to have the instruments on hand that are described here, because they will tell you different things and will find their best use in different situations.

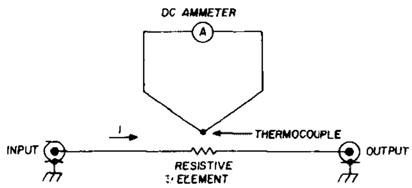



**fig. 1. Internal circuit of an rf ammeter uses a thermocouple to sense heating in a resistance element.**

### rf ammeters

Radio-frequency currents can't be measured easily by regular ac ammeters but require a special type of instrument based on the thermocouple. The internal schematic of an rf ammeter is shown in fig. 1. The rf current, $I$, flows through a resistive element. Power lost in this resistance is converted to heat, which is applied to a thermocouple. The thermocouple consists of two strips of dissimilar metals joined to form a V. When this junction is heated, dc will flow through the meter. This direct current will be proportional to the rms value of the rf current. Rf ammeters are generally free of frequency effects up to 50 or 60 MHz. Some instruments, especially older types, must be used with care since they may be sensitive to the type of panel on which they're mounted. Some, for example, are marked "calibrated on a 1/8-inch (3mm) steel panel" and are only accurate when used with such a panel. The rf ammeter can be used to indicate current levels in a transmission line or as a means of calibrating a more convenient form of rf wattmeter. The power-indicating instrument is connected in series with the rf ammeter and is inserted between the transmitter and a non-inductive dummy load. Various power levels are applied

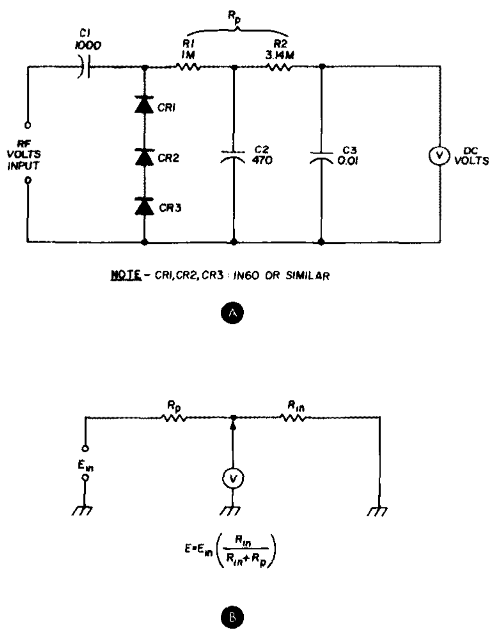


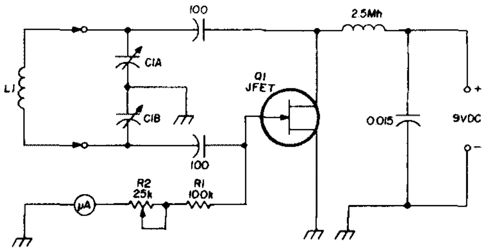

**fig. 2. Rf voltmeter adapter for a vtvm. Sketch (A) shows a typical probe; (B) shows a voltage divider for indicating rms values.**


**By Joseph J. Carr, K4IPV,** 5440 South 8th Road, Arlington, Virginia 22204

and the wattmeter is calibrated from the ammeter indications and the $P = I^2R$ relationship.

## rf voltmeters

Rf voltmeters can be made using a dc voltmeter and a special probe or adapter. Such probes are called "demodulator" probes. This probe is shown in **fig. 2A**. Diodes CR1-CR3 are standard germanium signal-detector devices such as the popular 1N60. Use one diode for every 15 or so volts of peak rf voltage amplitude. Capacitors C2 and C3 are filters. This type of probe is a




NOTE – C1-100pF/SECTION

**fig. 3. Circuit for a dip meter uses a jfet as the active element. This circuit is known as a gate dipper.**

peak-reading type, but most of our interest will be in the rms values. These values are related by:

$$E_{rms} = \frac{E_{peak}}{1.414} \tag{1}$$

Since a resistor is needed in the probe, and the voltmeter has an input impedance, we can construct a divider so that the voltmeter will read the rms value of rf sine waves. The voltage divider in **fig. 2B** follows the rule

$$E_{meter} = E_{in}\ \frac{R_{in}}{R_{in} + R_p} \tag{2}$$

Where

$R_{in}$ is the meter input resistance.
$R_p$ is the probe resistance.
$E_{in}$ is the input rf-signal peak voltage.

By combining the two equations we see that

$$\frac{R_{in} + R_p}{R_{in}} = 1.414 \tag{3}$$

But $R_{in}$ is usually equal to 10 megohms in most electronic voltmeters, so

$$\frac{10\ meg + R_p}{10\ meg} = 1.414 \tag{4}$$

$$R_p = (1.414)(10\ meg)\ or\ 14.14\ megohms$$

Use precision 1% (or better) noninductive resistors. The resistors in **fig. 2A** are shown as two separate components because they were available.

## dip meters

Originally known as grid-dip meters, these instruments are used to find resonant frequencies of tank circuits, or as a signal source, or to find (approximately) the values of capacitors and inductors. **Fig. 3** shows a typical dip-meter circuit; this one is a gate dipper using a junction fet. Inductor **L1** is mounted outside the instru-

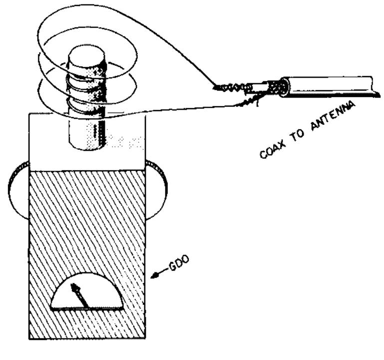
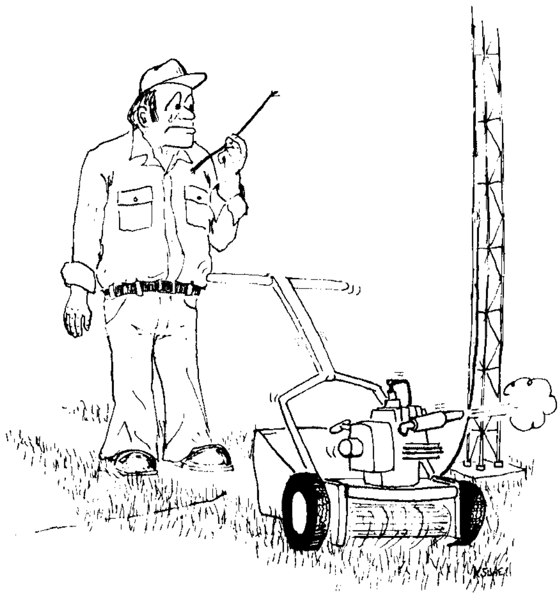



**fig. 4. Method for coupling a dip meter to an antenna.**

ment case. Energy from this coil is coupled to the coil in the resonant tank circuit being tested. When the L1-C1 tank of the dipper is tuned to the resonant frequency of the tank being tested, the amount of energy transferred will increase sharply which produces a slight dip in gate current. This property can be used to indicate the resonant frequency of the tank under test, which is read from the calibrated dial ganged to C1.

Dip meters usually are not too accurate as indicators of frequency, so a calibrated receiver should be used to sense the signal and read its frequency.

An antenna is basically a tuned resonant circuit, so it can be tested using the dip meter. **Fig. 4** shows a method by which the dipper is coupled to an antenna. The point at which the dip is noted is the antenna resonant frequency. A mistake often made when using dippers is



tuning too fast. The dip is slight, so one must tune very slowly.

## Wheatstone bridge

Basic to many forms of instrument is the Wheatstone bridge of fig. 5. This circuit compares the outputs of two voltage dividers Z1/Z3 and Z2/Z4. Under these circum-

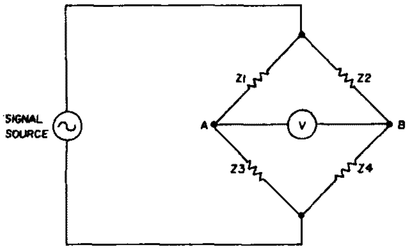



**fig. 5. Wheatstone bridge, which is the basis of many test instruments.**

stances, where Z1/Z3 = Z2/Z4, the voltage between points A and B is zero. Under these conditions,

$$Z4 = \frac{Z3Z2}{Z1} \tag{5}$$

In most cases impedance Z3 will be a calibrated standard resistance and Z4 will be the unknown. The values of Z1 and Z2 are not *too critical*, but they *should not be too* different from the other resistances in the circuit. This keeps the necessary amplitude of the signal source within reasonable bounds. In many cases $Z1 = Z2$ so that $Z3 = Z4$ at null. This makes calibration of Z3 easier. The meter at the center of the bridge may be either a voltmeter or a current meter. In fact, it is the current meter that's most frequently used as the null indicator.

## the noise bridge

An adaptation of the Wheatstone bridge is shown in fig. 6A. This is the noise bridge and uses a diode noise generator and amplifier to drive two arms of the bridge. Transformer T1 is a toroid with trifilar winding. One coil is connected across the output of the noise generator amplifier, while the remaining two coils form arms of the bridge circuit. In another arm of the bridge is a calibrated 0-250 ohm potentiometer, R1. The last bridge arm is the antenna or other unknown.

The null detector in this version of the bridge circuit is a receiver. Although the coax cable to the receiver may be any convenient length, the cable to the antenna must be either extremely short relative to a half wavelength or it must be an integer multiple of half wavelength (1, 2, 3, . . .). This length is found from

$$L = \frac{492}{f_{MHz}} \cdot (V \cdot N) \tag{6}$$

where $V$ is the cable velocity factor (0.8 for foam dielectric and 0.66 for regular coax) $N$ is any integer. This length is necessary because the impedance at the antenna feedpoint is reflected at half-wavelength intervals along the line. The noise bridge is connected to both the receiver and the antenna. The receiver is tuned to the desired frequency. Resistor R1 is then varied slowly until a null in noise level from the receiver is noted.

Fig. 6B shows an alternative form of noise bridge in which a 140-pF variable capacitor is connected in series with a 250-ohm potentiometer. A compensating 70-pF fixed capacitor is connected in series with the antenna. The variable capacitor is fitted with a calibrated dial that has a zero marked at the point where C1 = 70 pF (in other words, C1 = C2). Arbitrary calibration marks are scaled ± from this center zero. A null occurring when C1 is set to its zero point (70 pF) means that the antenna is resistive, which implies that the receiver is tuned to the antennas resonant frequency. If the null occurs either side of the zero point, then the antenna is reactive at the frequency indicated on the receiver.

The scale is marked $X_C$ on one side of the zero point and $X_L$ on the other side. If a null occurs on the $X_C$ side, then the receiver is tuned on the high side of resonance. Alternatively, if the null occurs on the $X_L$ side, then the receiver is tuned to a frequency that is too low. The antenna is resonant at the frequency indicated

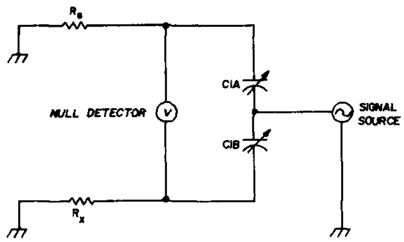



**fig. 7. The antenna impedance bridge, another adaptation of the Wheatstone circuit.**

on the receiver dial when a null occurs at a point where the capacitor dial reads zero (C1 = 70 pF). Amateur noise bridges are made by Omega-T Systems and by Palomar Engineers.

## antenna impedance meters

Another adaptation of the Wheatstone bridge is the antenna impedance bridge of fig. 7. Bridge arms consist

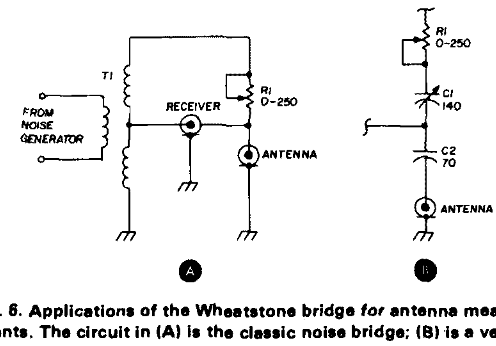



**fig. 6. Applications of the Wheatstone bridge for antenna measurements. The circuit in (A) is the classic noise bridge; (B) is a version in which a capacitor is used to null reactance.**

of C1A, $R_s$, C1B, and $R_x$. Capacitor C1A and C1B are respective halves of a differential capacitor. In this type of variable, capacitor C1A increases as C1B decreases. $R_s$ is a standard resistance, 50 or 75 ohms in most cases. $R_x$ is the resistive component of the antenna impedance.

This circuit will not measure reactance but it does allow you to tell whether there is a reactive component

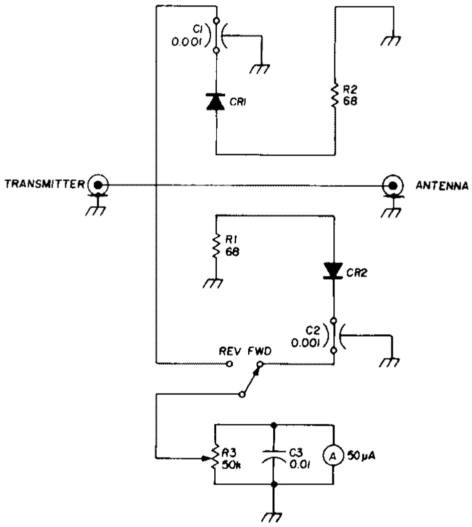



**fig. 9. The Monimatch swr mater, popularized by ARRL publications. Heart of the circuit is the pickup sensor, which consists of two conductors, one for forward and backward swr indication (see text).**

present. In a purely resistive situation the meter will null all the way to zero when the dial on C1 is set so that the bridge is balanced. If the meter does not null all the way to zero, then there is some reactance present. You can tell whether it is capacitive or inductive by varying the signal-source frequency until a frequency is found where the meter does null to zero. Antennas are capacitive below their resonant frequency and inductive above their resonant frequency. This information, then, tells you whether to lengthen or shorten the radiator to achieve resonance at the desired frequency.

Signal sources for the antenna impedance meter include ordinary signal generators, dip meters, or individual crystal oscillators. If you're very careful and use loose coupling, it's also possible to use a low-power transmitter as the signal source. I've used both a dip meter and those low-cost International Crystal OX oscillator kits in this application. I've also used the Leader LIM-870A antenna impedance meter (1.8 - 150 MHz). It is designed to mate with their dip meter, back-to-back, so that a complete signal source-bridge assembly may be formed.

## swr bridges

There are several ways to measure antenna swr. One is to measure the voltage components of the forward and reflected power at some point in the transmission line. This relationship is given by:

$$swr = \frac{V_f - V_r}{V_f + V_r} \qquad (7)$$

where:

$V_f$ is the forward voltage
$V_r$ is the reflected voltage

Voltage varies along the line, and it's possible to reduce the *apparent* swr by trimming the coaxial cable. However, this results in an erroneous reading that may give you a false sense of security.

**Fig.** 8 shows a resistive swr bridge. It is a Wheatstone bridge in which R1, R2, R3 and the antenna impedance form the respective arms. Diodes CR1 and CR2 should be a matched pair but should be satisfactory if you match their forward and reverse resistances with an ohmmeter. Resistor R5 is used to trim out differences in R6 and R7. Normally, the microammeter is set to full scale with S1 in the forward position. Resistor R4 is adjusted to make M1 indicate full scale (100 $\mu$A). The switch is then turned to the reflected position and the meter needle drops to a current proportional to the swr. In most swr meters the dial will be calibrated in swr units.

Another type of swr indicator popularized in certain ARRL publications, is the Monimatch of **fig. 9**. The heart of this instrument is the pickup sensor. It consists of two conductors, one for each direction, arranged in parallel and in close proximity to the coaxial cable center conductor. In some versions enameled wires are slipped underneath the shield conductor of a short

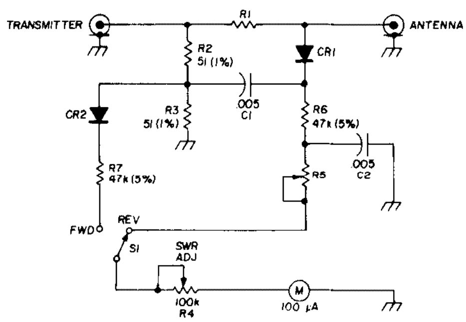



**fig. 8. Resistive vswr bridge. R1 is 51 or 75 ohms, depending on coax-cable impedance.**

length of coaxial cable. In others, the conductors are etched onto a piece of printed-circuit board inside a shield box. This class of instrument is different from other swr meters in that it becomes more sensitive as frequency increases.*

## rf wattmeters

**Figs. 10** and **11** show two popular forms of rf power meter. Of course, you can compute the swr of an antenna system by comparing forward and reflected power levels. The circuit of fig. 10A is known as the Micro-

*For still another swr instrument, see "Using the swr Indicator," *repair bench, ham radio*, January, 1977, page 66. **Editor**

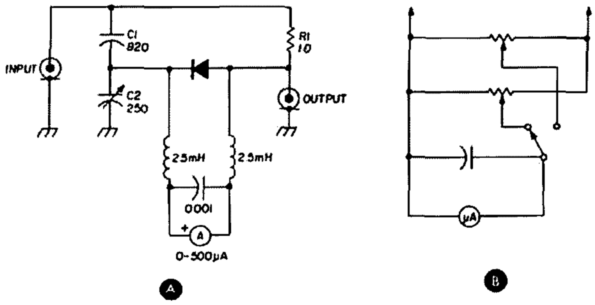



**fig. 10. The Monimatch circuit for an rf wattmeter. (A). Various power ranges can be selected by the circuit in (B).**

match, which was available commercially for years. It is also based on the ubiquitous Wheatstone bridge.

The bridge arms consist of C1, C2, R1 and the impedance of the antenna. R1 is only 1 ohm, so very little power is lost due to insertion of the instrument into the line. Various versions of the Micromatch have one, two, or three power ranges (10, 100, or 1000 watts full scale). These are selected as in fig. 10B.

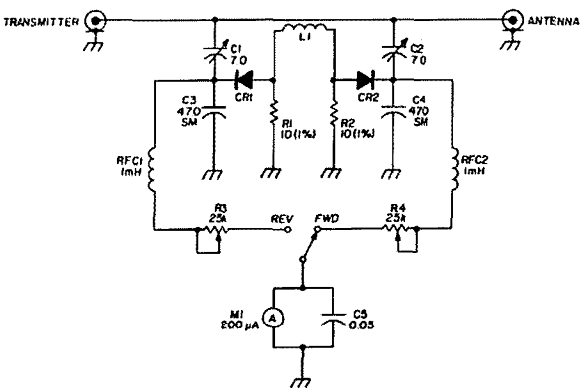



**fig. 11. Relfectometer principle for an rf wattmeter. Pickup sensor makes this instrument unique.**

A more recent circuit based on the reflectometer principle is shown in fig. 11. Although much of this circuit is similar to other circuits, we again have a situation in which uniqueness is in the design of the pickup sensor. In this case it consists of a coil of wire wrapped around the coaxial cable center conductor. In most such instruments the conductor is actually a heavy piece of solid bus wire and the pickup coil is wound on a toroid form. The wire conductor is threaded through the center of the toroid. This, in fact, is the basis of most low cost rf wattmeters on the market today.

Although it's a little excessive to expect any one amateur to own all of these different types of antenna instrumentation, it must be noted that every one of them is unique in its own way and all are eminently useful for doping out or troubleshooting reluctant antenna systems.

ham radio

# sync generator
# for a black-and-white 525-line television system

## The National MM5320 IC is featured in this inexpensive circuit

**This article describes** a sync generator that supplies horizontal and vertical drive, blanking, and sync signals for a standard black and white interlaced 525-line TV system. It is simple and easy to build on the PC boards available,* or it can be reduced in size and mounted inside your TV camera. Almost all functions are performed within a single IC, the National MM5320 (about $5).

The PC board, loaded with parts, weighs in at only 4 ounces (113g). However, this lightweight board contains the equivalent of a 40 tube sync generator of the past and requires no maintenance to keep it running. Power consumption is only a few watts.

### circuit description

The schematic is shown in **fig. 1**. The master oscillator consists of a 2.04545-MHz crystal, Y1, and a Motorola MC4024P multivibrator IC, U1. Actually, two multivibrators are contained in U1, so the spare is available at pin 12 on the PC board, **fig. 2**.

The circuit works well and delivers a symmetrical square wave at TTL level with almost any crystal. A disadvantage is that no provision is included for pulling the crystal to exact frequency. It's not necessary to do this with this sync generator — just plug in any HC6U crystal with the correct frequency.

The MM5320 has more features than needed for this application, such as gen-locking capability with circuits to reset both horizontal and vertical counters and a field-index output identifying the odd field at a 30-Hz rate. The MC5320 can be programmed to operate with either a 1.26-or 2.04545-MHz crystal. Therefore, certain pins are jumpered for this application, as shown in the PC-board layout, **fig. 2**.

The MM5320 also has burst-flag pulses, so it has color capability if a few circuits are added. Basically, a color generator would have a stable crystal-controlled master oscillator operating at 14.31818 MHz. This frequency is divided twice to derive the 3.58-MHz color subcarrier, which is processed to obtain a 2 volt p-p sine wave with adjustable phase shift.

The master oscillator also could drive a 7:1 counter to obtain 2.045 MHz, which is needed to drive the MM5320. The specification sheet for the MM5320 shows a minimum pulse width requirement of 190 nanoseconds. The 7:1 divider output is a 70-nanosecond pulse, so some means, such as a one-shot, could be used to square up the wave before it is applied to the MM5320. The master oscillator should remain within 4.0 Hz of 14.31818 MHz, so it can be seen that a few small problems need to be worked out. For black-and-white TV, we can ignore these difficulties and use the 2.04545-MHz crystal.

The circuit is simple. Buffering should be used between the mos-generator chip and output loads to

*An undrilled epoxy board is available from Bert Kelley, 2307 South Clark Avenue, Tampa, Florida 33609, for $6.00 postpaid in U.S.A.

**By Bert Kelley, K4EEU**, 2307 South Clark Avenue, Tampa, Florida 33609

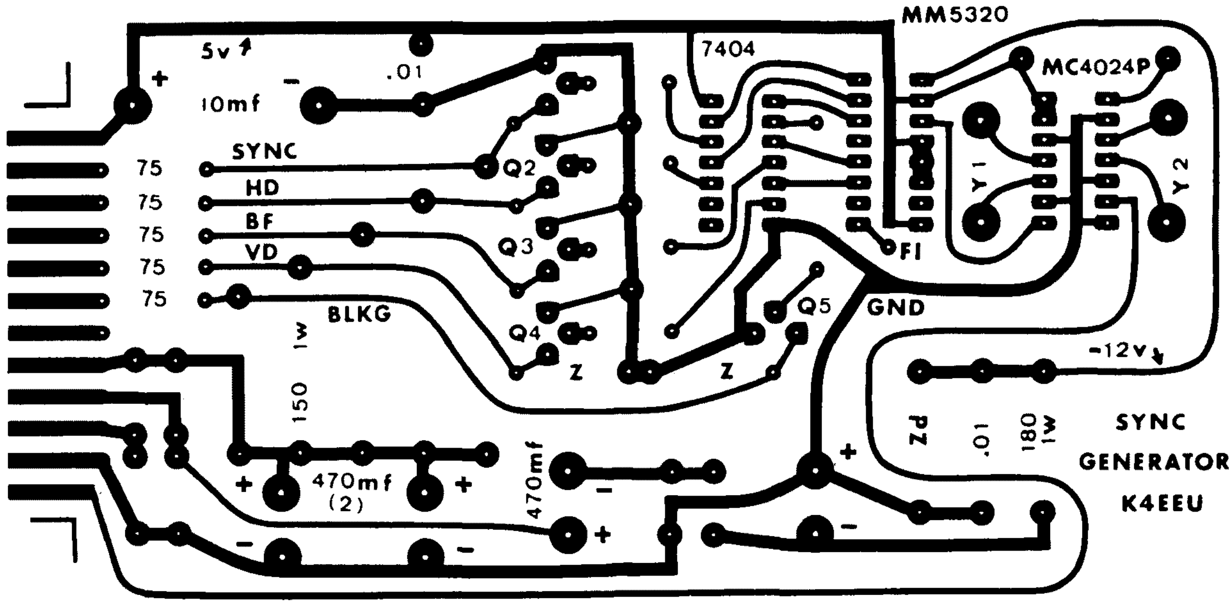



fig. 2. PC-board layout for the TV sync generator. Boards are available for $6.00 postpaid in U.S.A. from the author.

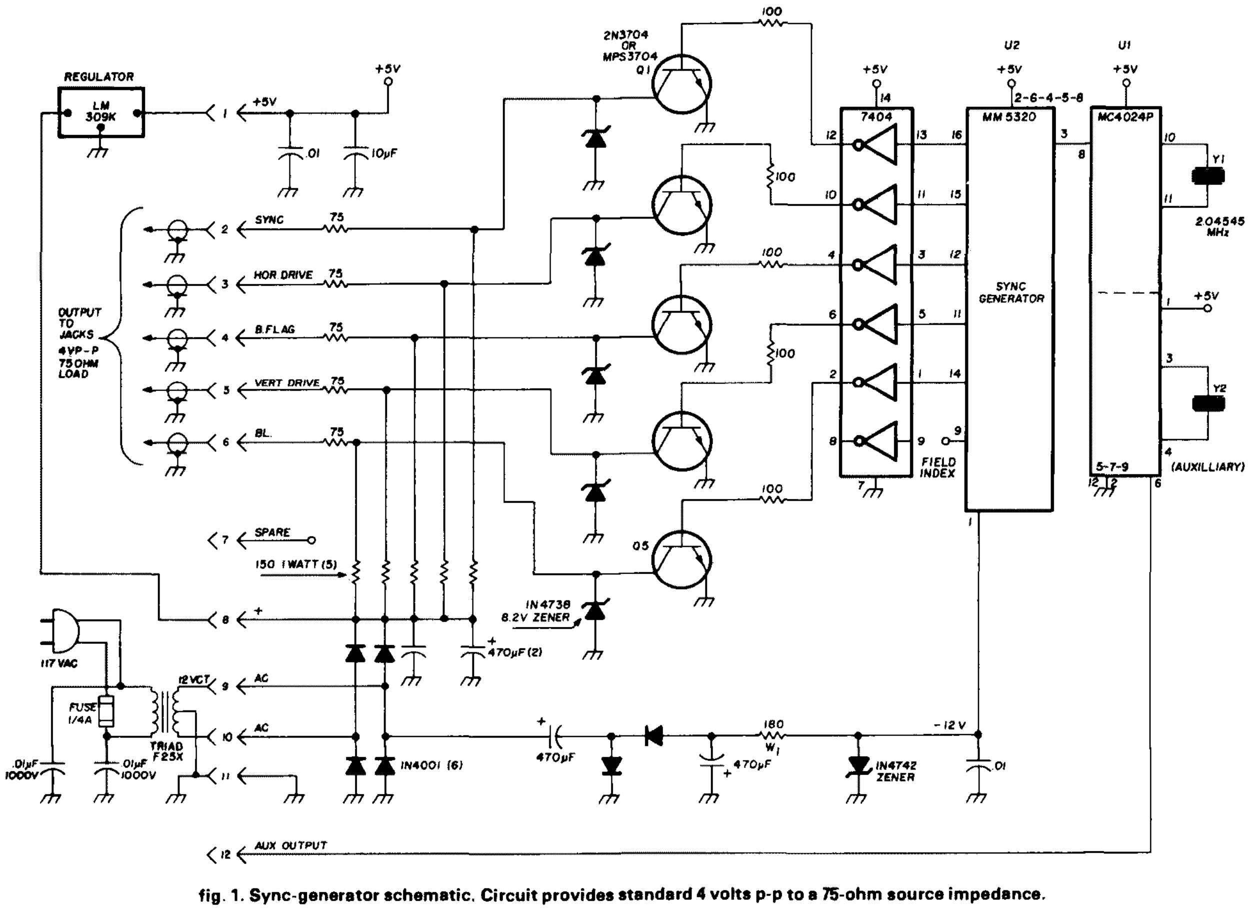
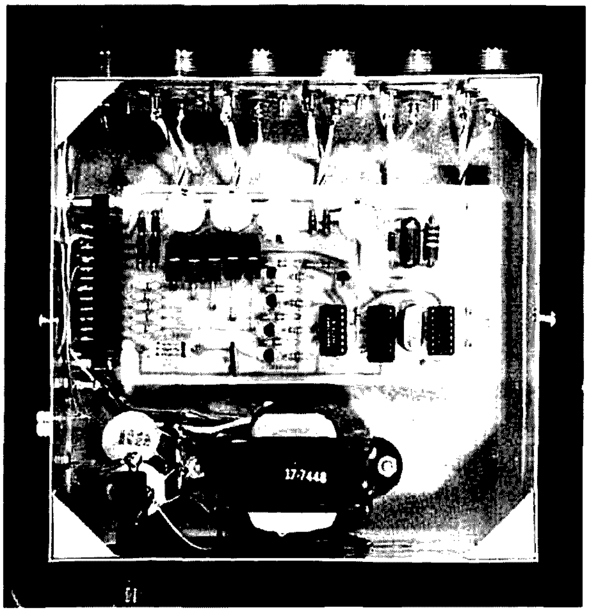



fig. 1. Sync-generator schematic. Circuit provides standard 4 volts p-p to a 75-ohm source impedance.

keep transients from damaging the chip. The circuit shown provides a 4-volt p-p to a 75-ohm load at 75-ohm source impedance. The voltage at the transistor collectors is limited by 8.2V zener diodes; when the transistor saturation voltage is deducted, the p-p voltage swing is almost exactly 8 volts to give 4 volts under load (negative-going pulses).

## construction

The circuit board, which is 3 by 6 inches (77 by 153mm), fits an Amphenol 142-012-01 12-pin connector. (Alternatively, connections can be soldered to the output pins.) IC sockets should be used, at least for the mos chip. Use 150-ohm 1-watt resistors where indicated. The 470-$\mu$F filter capacitors are of the radial-lead type. I used SO-239 output connectors; however, ordinary phono connectors would be all right (and much cheaper).




**Underchassis view of the TV sync generator showing power supply, right; PC board, canter; and peripheral wiring. An LMB chassis is used, which also provides a heat sink for the regulator circuit.**

The circuit board should be mounted in a shielded enclosure such as the LMB 772, which also serves as a heat sink for the LM309K regulator.

The MM5320 is available from Nexus, and other parts are available from James Electronics.*

**ham radio**

*Nexus Company, P.O. Box 3357, San Leandro, California 94578 and James Electronics, 1021 Howard Avenue, San Carlos, California 94070.

MICROPROCESSORS

# microcomputer interfacing: logical instructions

Most microcomputers manipulate information eight bits at a time. For example, the 8080A chip can move eight bits from internal register to internal register, between the accumulator and an external I/O device, and from internal register to memory. In additon, it can also perform arithmetic and logical operations, including add, subtract, and compare, and logical instructions AND, OR, Exclusive-OR, and complement. In this column, we will be concerned with the logical operations.

A truth table is the basic form that governs the one-bit logic operations. As a tabulation, it shows the relationship between all possible combinations of input logic levels and the subsequent outputs in a manner that completely characterizes the circuit functions.[1]

The truth tables for the AND, OR, Exclusive-OR, and complement operations are:

| AND | | | OR | | | Exclusive-OR | | | complement | |
|---|---|---|---|---|---|---|---|---|---|---|
| B | A | Q | B | A | Q | B | A | Q | A | Q |
| 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 1 |
| 0 | 1 | 0 | 0 | 1 | 1 | 0 | 1 | 1 | 1 | 0 |
| 1 | 0 | 0 | 1 | 0 | 1 | 1 | 0 | 1 | | |
| 1 | 1 | 1 | 1 | 1 | 1 | 1 | 1 | 0 | | |

These truth tables are called *one-bit tables* because the data words, A and B, each contain only a single bit.

When discussing logic instructions, it is useful to employ *Boolean symbols*. Such symbols originate from the subject of *Boolean algebra*, which is the mathematics of logic systems. This particular form of mathematics was originated in England by George Boole in 1847. Alphabetic symbols such as A, B, C, . . ., Q are used to represent logical variables with 1 and 0 representing logic states. Boolean algebra did not become widely used until 1938, when Claude Shannon adapted it to analyze multi-contact networks *for* telephone networks.

You should learn the basic Boolean symbols that are used in Boolean algebra computations, and thus all digital logic. These symbols include the following:

- $+$ which means *logical addition* and is given the name OR
- $\bullet$ which means *logical multiplication* and is given the name AND
- $\oplus$ which is given the name *Exclusive-OR* or XOR
- $\overline{A}$ which means *negation* and is given the name NOT

The negation symbol is a solid bar over a logical variable. Thus, the Boolean statement for a 2-input AND gate is $Q = A \bullet B$, or simply $Q = AB$, where the equality symbol ( = ) means that both variables or groups of variables are the same. It is useful to summarize the symbol operations for the three gates that are being considered:

| AND | OR | Exclusive-OR | complement |
|---|---|---|---|
| $0 \bullet 0 = 0$ | $0+0=0$ | $0+0=0$ | $\overline{0}=1$ |
| $0 \bullet 1 = 0$ | $0+1=1$ | $0+1=1$ | $\overline{1}=0$ |
| $1 \bullet 0 = 0$ | $1+0=1$ | $1+0=1$ | |
| $1 \bullet 1 = 1$ | $1+1=1$ | $1+1=0$ | |

Multi-bit logic operations are treated as many one-bit logic operations, therefore no new principles of logic are involved. The corresponding bits of one binary word logically operate on the corresponding bits of the second binary word to produce an overall multi-bit logic result. The length of the binary words can be any number of bits: two bits, eight bits, thirty-two bits, etc. Since the 8080A microprocessor performs multibit logic operations on eight-bit words, all of our examples will involve full bytes.

Consider the eight-bit logic variable, A. The individual bits in this variable can be labeled as A7, A6, A5, A4, A3, A2, A1, and A0, with A0 being the least significant bit ($2^0$ bit) and A7 being the most significant bit ($2^7$ bit). Also, consider the eight-bit logic

**By Jonathan Titus, David G. Larsen, WB4HYJ, and Peter R. Rony**

Mr. Larsen, Department of Chemistry, and Dr. Rony, Department of Chemical Engineering, are with the Virginia Polytechnic Institute and State University, Blacksburg, Virginia. Mr. Jonathan Titus is President of Tychon, Inc., Blacksburg, Virginia.

variable, B, which has individual bits that can be labeled as B7, B6, B5, B4, B3, B2, B1, and B0. The logic operation, $A \bullet B = Q$, means the following eight one-bit logic operations:

| | |
|---|---|
| $A0 \bullet B0 = Q0$ | $A4 \bullet B4 = Q4$ |
| $A1 \bullet B1 = Q1$ | $A5 \bullet B5 = Q5$ |
| $A2 \bullet B2 = Q2$ | $A6 \bullet B6 = Q6$ |
| $A3 \bullet B3 = Q3$ | $A7 \bullet B7 = Q7$ |

The result of the logic operation is the logic variable, Q, which has a least significant bit of Q0 and a most significant bit of Q7. In other words, *multi-bit logic operations are performed bit by bit via a series of one-bit logic operations*. It is easier to perform multi-bit logic operations if the multi-bit binary words are placed one under the other. Thus, if $A = 11011111_2$ and $B = 00100011_2$, then $A \bullet B$ is

$$\begin{array}{r} 11011111_2 \\ \bullet 00100011_2 \\ \hline 00000011_2 \end{array}$$

or $Q = 00000011_2$. We have performed a logical AND, and have used the relationships $0 \bullet 1 = 0$ and $1 \bullet 1 = 1$ in deriving the final result.

One of the more important uses for multi-bit logic operations is in situations in which the on-off state of external devices must be monitored. Consider the following system of eight devices:

| bit position | devices | logic stats information |
|---|---|---|
| Bit 0 | pressure sensor | 1 = pressure above setpoint<br>0 = pressure at or below setpoint |
| Bit 1 | temperature sensor | 1 = temperature above setpoint<br>0 = temperature at or below setpoint |
| Bit 2 | velocity sensor | 1 = velocity above setpoint<br>0 = velocity at or below setpoint |
| Bit 3 | flow rate sensor | 1 = flow rate above setpoint<br>0 = flow rate at or below setpoint |
| Bit 4 | concentration sensor | 1 = concentration above setpoint<br>0 = concentration at or below setpoint |
| Bit 5 | valve A | 1 = valve A open<br>0 = valve A closed |
| Bit 6 | valve B | 1 = valve B open<br>0 = valve B closed |
| Bit 7 | power | 1 = power on<br>0 = power off |

We can call the group of eight bits the *status byte* for our system of eight devices. At any instant of time, the status byte will have a specific value. For example, the status byte $11100010_2$ signifies that the pressure is at or below the setpoint, the temperature is above the setpoint, the velocity is at or below the setpoint, etc.

A group of logical instructions will permit you to determine the following characteristics about the previously listed external devices.

1. Which devices are on, open, or above the setpoint?
2. Which devices are off, closed, or at or below the setpoint?
3. Since the last time we checked, which devices have gone from on to off, open to closed, or above the setpoint to at or below the setpoint?
4. Since the last time we checked, which devices have gone from off to on, closed to open, or at or below the setpoint to above the setpoint?

In other words, using logical instructions you can determine not only the current state of the external devices, but also what changes have occurred since the last time that the devices were interrogated. Assume that we have just interrogated all eight devices and have found the current status byte to be $11101010_2$, where the least significant bit, bit 0, is on the far right. One second ago, the status byte was $11101001_2$. We wish to know what the current state of each device is, which devices have changed state during the last second, and in which direction. The steps that are employed to answer the questions are as follows:

1. Examine the current status byte. Determine the status of each external device from the logic state of its status bit.

The current status byte (CSB) is $11101010_2$. From this value, we conclude that the pressure, velocity, and concentration sensors are all at or below their respective setpoints; the temperature and flow rate sensors are above their respective setpoints; and that value A, value B, and power are all on.

2. Perform an Exclusive-OR operation between the prior status byte (PSB) and the current status byte (CSB). A logic 1 in the result will indicate that the logic state of that device has changed.

The logical operation that we wish to perform is,

$$PSB \oplus CSB = Q1$$

where $PSB = 11101001_2$, $CSB = 11101010_2$, and Q1 is the result of the Exclusive-OR operation. Thus,

$$\begin{array}{rl} 11101001 & \text{PSB} \\ \underline{\oplus 11101010} & \text{CSB} \\ 00000011 & \text{Q1} \end{array}$$

and $Q1 = 00000011_2$. We conclude that only the pressure and temperature sensors have changed state.

3. Perform an AND operation between Q1 and the prior status byte (PSB). A logic 1 in the result indicates a device that has changed state from logic 1 to logic 0.

The logical operation that we wish to perform is,

$$PSB \bullet Q1 = Q2$$

where $PSB = 11101001_2$, $Q1 = 00000011_2$, and Q2 is the result of the AND operaton.

| | |
|---|---|
| 11101001 | PSB |
| •00000011 | Q1 |
| 00000001 | Q2 |

Thus we can conclude that the pressure sensor has changed from being above the setpoint to now at or below the setpoint (logic 1 to logic 0 transition).

4. Negate (or complement) Q2, then ADD this complemented result with Q1. A logic 1 in the result indicates a device that has changed state from logic 0 to logic 1.

The logical operation that we now perform is,

$$\overline{Q2} \bullet Q1 = Q3$$

Since Q2 is $00000001_2$, the complemented value of Q2 must be $11111110_2$. The result of the AND operation is obtained as follows:

| | |
|---|---|
| 11111110 | $\overline{Q2}$ |
| •00000011 | Q1 |
| 00000010 | Q3 |

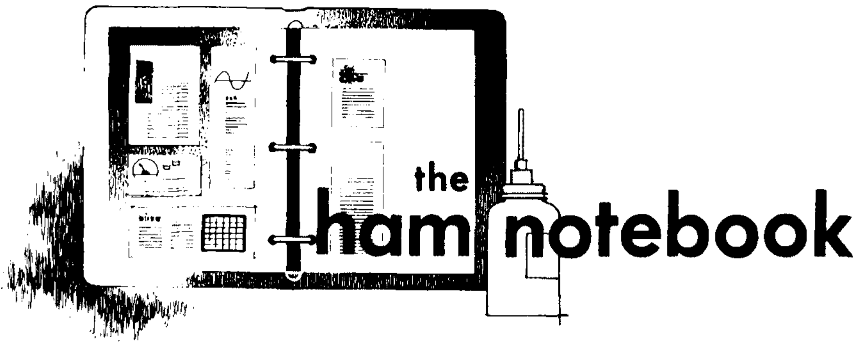
The result, $Q3 = 00000010_2$, permits us to conclude that the temperature sensor has changed from at or below the setpoint to above the setpoint (logic 0 to logic 1 transition).

The reason that we perform a series of logical operations is to determine what type of corrective actions must be applied to our system if it is not operating properly. If the temperature is below its setpoint, we may have to turn on a heater. If the concentration of the reactant is too high, we may have to turn valve B off and temporarily halt the flow of reactant into the system. If the pressure is above the setpoint, we may have an emergency condition and must shut the power off to the entire system. With a properly interfaced microcomputer, all such decisions can be easily and quickly made under software control and the necessary corrective actions initiated. However, you must be aware of the fact that mechanical and electromechanical devices typically have response times that are much longer than the decision times for a microcomputer. These response times must be taken into account in our software design, and are important considerations in the field of digital controls.

## reference

1. R. F. Graf, *Modern Dictionary of Electronics*, Howard W. Sams & Co., Indianapolis, 1972.

**ham radio**




## low-impedance headphones for Heath HW-16

The Heathkit HW-16 transceiver represents a very good value in amateur radio equipment, offering the beginning ham a transmitter and receiver combination designed for CW, and contains everything necessary to "get on the air" except antenna, speaker and key. The HW-16 also makes a good back-up rig in case your main station equipment is out of service or if your General class license is running out and you need some CW contacts to brush up on your code speed and fulfill the renewal requirement.

The problem arises when trying to use 8-ohm headphones to avoid disturbing the rest of the family. Low-impedance headphones do not mute the speaker. Fortunately, the solution is simple, quick, and costs nothing. If you are building the kit, you might want to wire in the new circuit from the start.

**Fig. 1A** is a reproduction of the audio output circuit diagram which appears in the Heath HW-16 construction and operation manual.

You will see that plugging headphones into this circuit does not disconnect the speaker, but places the speaker in series with the phones. With 8-ohm phones and an 8-ohm speaker, one-half of the audio still goes to the speaker. You will also note that to use earphones of any kind, you must have an 8-ohm load plugged into the speaker jack, because the two jacks are in series.

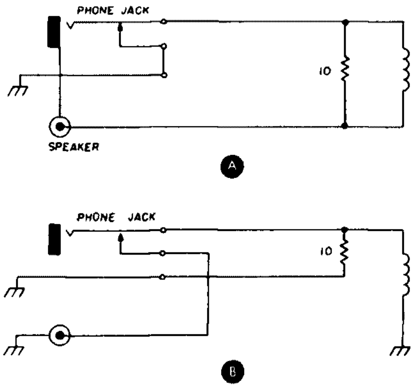



**fig. 1. Schematic of the audio output circuit of the Heathkit HW-16 transceiver showing phone jack in series with the speaker jack as originally wired. (B). Revised schematic showing simple modification of the HW-16 audio circuit permitting phones or speaker to be used interchangeably.**

Rewire the circuit as shown in **fig. 1B**, to disconnect the speaker completely when phones are plugged in. Note that in this circuit, however, you can use phones *or* speaker interchangeably. Since I've rewired my HW-16 this way, I operate more often because my family doesn't complain that audible CW notes interfere with their favorite television program.

**John Pawlicki, WN8WJR**

## improved ssb reception with the Collins R392

A serious deficiency of the Collins R392, an otherwise excellent general-coverage receiver, is the severe distortion produced by strong ssb signals. While turning down the rf gain control solves the problem, it is extremely inconvenient when dealing with a variety of signals of different strengths — as when tuning across an amateur band.

The R392 was designed primarily for a-m and fsk use and uses a simple diode detector. Experiments with a product detector and outboard audio stages produced no significant improvement. Observing the signals at the i-f output connector on the front panel with an oscilloscope revealed that strong ssb signals caused considerable flat-topping in the stages ahead of the detector. When the rf gain control was backed off to the point where clipping did not

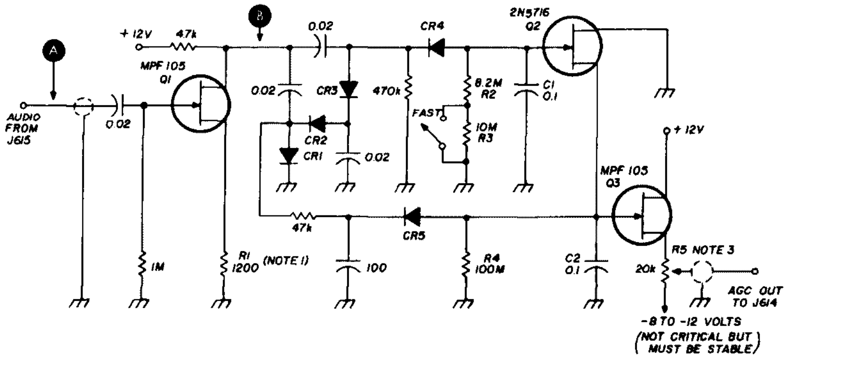



**fig. 2. Hang agc circuit for the Collins R392 communications receiver. R4 is five 20-megohm resistors in series. Resistor R1 is adjusted for 7 Vdc at rest point B. Adjust R5 for zero out at agc out with no signal at the input (point A). All diodes are 1N459A or 1N914.**

occur, the detected audio sounded fine. This indicated that the problem was due to the agc system (designed for a-m signals) more so than the detector.

Ssb reception may be improved using the audio-derived hang agc circuit of **fig. 2.** The detector/audio module of the R392 conveniently has test points at the detector output, at which the audio signal may be picked up, and at the agc line, at which the externally-derived agc voltage may be introduced, with no modifications.

A hang-type circuit derived from that described by Hartke* is used. Q1 provides some amplification and high input impedance to prevent loading of the audio circuit at J615. The output of Q1 is rectified by CR1 and gated by CR5 onto C2, the agc hang capacitor. Simultaneously, CR1, CR2, CR3, and CR4, in a voltage quadrupler configuration, develop a control voltage gated to C1 via CR4. This control voltage keeps Q2 pinched off as long as a signal is present and permits C2 to remain charged and maintain the negative agc voltage. Upon removal of the signal, C1 discharges through R2/R3 and – after the hang time, when the pinch off voltage of Q2 is reached – it conducts and quickly discharges C2, restoring the receiver gain.

Source follower Q3 acts as a high input impedance buffer between C2 and the receiver agc line. R4 is required to provide a dc return for the gate of Q3. The time constant R4C2 is so long relative to R2C1 that the presence of R4 causes no significant discharge of C2 during the hang period. R5 serves to adjust the dc offset at the agc terminal. The minus supply voltage is not critical, as long as it is stable (I happen to have an 8-volt zener on hand). Strong signals charge C2 negatively, reducing the drain current of Q3 and causing the source voltage (drop across R5) to move in a negative direction.

### installation

No major modification of the R392 is required. The agc circuitry was built on a small piece of perforated board and mounted in the external power supply unit. Connections to the detector output and the agc line were made at test points J615 and J614 respectively, on the audio module (lower deck assembly). In my case, lengths of miniature shielded cable were fitted with pin plugs to match the J614 and J615 sockets and the other ends wired to unused pins on the receiver power socket, J103. The connections to the agc module in the power supply were thus made via the power connectors, again using miniature shielded cable. Pin J on the receiver power connector is unused and is directly available. Several other pins are only needed when the original matching transmitter and microphone/headset are used. Therefore, the wire to pin C was unsoldered, capped, and this pin on J103 was used for the other agc connection.

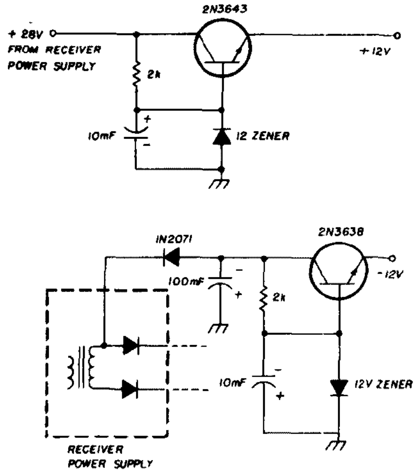



**fig. 3. Power supply for the R392 hang agc system (fig. 1).**

The front panel must be removed to gain access to the rear of the power connector. However, this is a relatively simple procedure following the instructions in the manual.† With the panel off, capacitor C104, 0.47 mF, which is switched into the original agc circuit by the bfo switch, should be disconnected. This improves the attack time of the new agc circuit. This is the larger of the two capacitors mounted on the inside of the front panel near the bfo on-off switch.

The required voltages are derived from the power supply unit shown in **fig. 3.** Simple zener diode regulators would also suffice since the current demands are only a couple of milliamps.

### adjustment

Adjustment is simple. First, with no input signal to Q1, source resistor R1 is selected to provide a static drain voltage of about 7 volts to ensure maximum dynamic range at this stage. Next, with the connection to J614 also open, R5 is adjusted for zero volts at the agc output terminal. The circuit is then ready to go.

The added agc required no butchering of the original circuitry and has resulted in a dramatic improvement in ssb reception performance.

**A.D. Lightstone, VE3LF**

*J.L. Hartke, W1ERJ, "Solid-State Hang AGC Circuit for SSB and CW," *ham radio*, September, 1972, page 50.

†"Field and Depot Maintenance Manual, Radio Receiver R-392/URR," *Army Technical Manual TM 11-5820-334-35.*

## test probe accessory

A volt-ohm-milliammeter (VOM) may be the most-used piece of equipment owned by the average radio amateur or experimenter. All voms are supplied with flexible insulated wire leads that have either pointed test probes or alligator clips at their ends, but not both.

Each type has its preferential use, but unfortunately not always in a compatible manner. Pointed test probes are handy for poking around to check resistors, capacitors, shorts, and the like; but for troubleshooting and checking voltages at tube sockets and so on, I prefer an alligator clip on the negative lead so that it may be clipped on to the chassis. Alligator clips are also handy for connecting to a circuit that may require monitoring for a period of time.

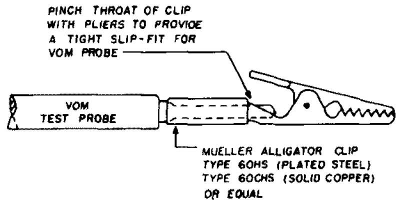


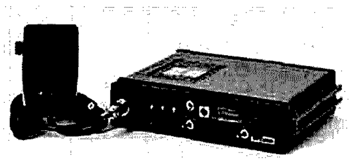

**fig. 4. Pointed-tip test probe is easily converted to clip-type test probe by simple addition of an alligator clip.**

My favorite vom had pointed test probes and, due to my experimental endeavors, I fashioned a bunch of patch cords with alligator clips on both ends. A few of these in red and black were left over, so I began experimenting with them. I found that I could slip the pointed test probes into the clips (Mueller) and obtain a reasonable fit. Now I have the advantage of using my test probes as is, or I can slip an alligator clip on when I want, as shown in **fig. 4.**

**Ken Cornell, W2IMB**

# NEW products

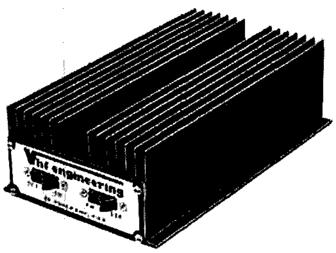
## synthesized two-meter fm



"The radio that goes where you do" is a feature of Wilson Electronics' new WE800 radio. The 800-channel, 2-meter portable synthesized radio has an internal nicad battery pack which allows installation as a mobile in the automobile or portable with shoulder strap and rubber duck. The unit provides 12 watts output in service, or 2 watt output as a portable with the internal nicad battery pack or ten AA nicads).

The WE800 has a low drain cmos synthesizer, with 45 milliamps on receive, and 450 milliamps on 2-watt transmit. Frequency range is 144-148 MHz in 5-kHz steps, 600 kHz offset up or down. Subaudible tones are available as an option. For more information, contact Wilson Electronics Corporation, 4288 S. Polaris, Las Vegas, Nevada 89103.

## 80-watt 450-MHz amplifier



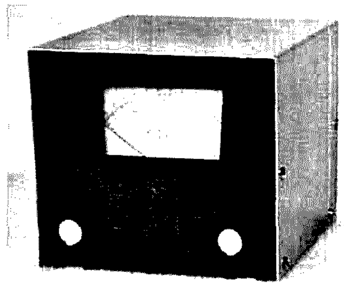
A new broadband, high efficiency four-mode amplifier for the 420-450-MHz amateur band has been introduced by VHF Engineering. This new amplifier, the Blue Line BLE 10/80, will deliver 80 watts output in either class C or linear mode with a nominal 20 watts input. It is designed to be used with fm, a-m, ssb, or CW rigs in the 10-watt class. A similar amplifier, the Blue Line BLE 30/80, is designed to be used as an amplifier for rigs in the 30-watt class.

The Blue Line series of amplifiers from VHF Engineering are high efficiency, broadband, stripline amplifiers which have been designed for long life and reliable operation. Because of their unique, broadband design, they contain no tunable or adjustable components. Tuning or adjustment will not be required during the lifetime of the units. Automatic transmit/receive switching is provided through the use of sensing circuitry which detects the modes. They are designed for 12-14 Vdc operation in base station or mobile service.

These new four-mode Blue Line amplifiers are manufactured by VHF Engineering, 320 Water St., Binghamton, New York 13902. The BLE 10/80 is priced at $289.95 and the BLE 30/80 is $259.95. Both units are wired and tested.

## antenna rotor control conversion kit

Have you had rotator damage? Removed the rotator? Waited for parts? No more! *Autobrak* is a complete conversion kit, including punched and finished cabinet for all HAM-M series 1, 2, and 3, rotator control units. *Autobrak* reduces the inherent problem of damaged rotator components due to instant brake engagement; it allows the antenna array to come to a coasting stop before brake engagement, thereby reducing stress on rotator components.



Other features include zener-regulated meter circuitry, adjustable brake delay, and handsome up-to-date styling compatible to most ham gear. The *Autobrak* is priced at $39.95 (plus shipping and handling, $1.75 in the United States). For more information write to Kampp Electronics, Post Office Box 43, Wheaton, Illinois 60187.

## vhf frequency counter

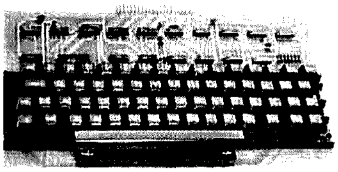

Sencore, manufacturers of high quality test equipment for radio-TV service and communications, is offering a new FC45 frequency counter with continuous frequency checking capability from audio through vhf (and uhf bands with an optional PR47 uhf prescaler).

The FC45 audio through vhf counter is highly sensitive, with 25 millivolts average throughout the band, for circuit servicing with a pickup loop that does


More Details? CHECK—OFF Page 126

not upset the circuit during frequency tests. The FC45 is highly accurate with a one part in one million tolerance, using a high-accuracy temperature-controlled oven. Accuracy is better than FCC requirements on all bands including uhf. An eight-digit display with all pushbutton action makes the FC45 extremely easy to use. An exclusive crystal checker is provided as an integral part of the FC45. The company has stated that the price of $395 is also far below any other vhf frequency counter on the market today; the optional PR47 prescaler, selling for only $125, is simply connected into the input cable and the FC45 readings multiplied by ten. If used with any other frequency counter, a separate PA202 Power Adapter at $9.95 is used. For more information write to Sencore, Inc., 3200 Sencore Drive, Sioux Falls, South Dakota 57107.

## ASCII keyboard encoder



A new ASCII Keyboard Encoder is now available from Radio Shack in their popular Archer Project-Board kit form. With the Project–Board concept you purchase the printed-circuit board with a complete, step-by-step assembly instruction manual and parts list. All necessary parts for the assembly of the ASCII Keyboard Encoder, including a 63-key computer control keyboard, are available from Radio Shack, or the builder may use parts from his own junkbox.

The completed keyboard encoder can be used to provide inputs to all types of equipment designed to operate with ASCII inputs, such as TV typewriters, minicomputers, microprocessors, electric typewriters, or any other devices which require positive or negative ASCII encoded alpha-numerical characters.

Features of the Archer ASCII Keyboard Encoder include a repeat key to control all characters and symbols, a

negative- or positive-going data valid strobe, latch outputs (stores last key code), shift and shift lock capability, true or false ASCII outputs, and six extra control keys. The encoder is able to handle an output of 833 characters per minute (cpm) and has a repeat key rate of 208 cpm. An external power source of 5 Vdc at about 500 mA is required to power the encoder.

The Archer ASCII Keyboard Encoder Project-Board with complete assembly and parts manual is priced at $14.95. All parts needed to assemble the encoder, including the circuit board, manual and keyboard, but excluding hardware and case, are available from Radio Shack for $57.80. Archer Project-Boards are available exclusively from Radio Shack stores and dealers in all 50 states and Canada.

## tape antennas

Tualatin Valley Labs is presently manufacturing indoor antennas for two meters. Available is a three-element beam at $9.95, and a vertical, ½-wave J pole design for $7.95.

The antenna elements are made from ¾ inch (2cm) wide aluminum tape designed specifically for rf applications. The adhesive backing on the tape is also conductive, allowing antenna connections to be made without the use of solder. Installation can be on any flat, dry surface, such as walls, doors, or inside closets. The tape alone is also available for your own experimentation in 10-foot (3m) lengths for $7.50 per 10 feet; 25 feet and 50 feet lengths are available for $15.00 and $25.00 respectively. For more information, write to Pat Adamosky, Tualatin Valley Labs, 18285 N.W. Parkview, Portland, Oregon 97229.

## receive multi coupler

Did you ever wish you had a way to use one antenna to feed two vhf receivers? Well, it's here! The *P13 Receive Multicoupler Unit*, from Hamtronics, Inc., is a dual-output fet preamplifier which provides low-noise gain and power splitting to drive two receiver inputs from one antenna.

The unit gives about 15 dB of gain in each of the two channels. If desired,



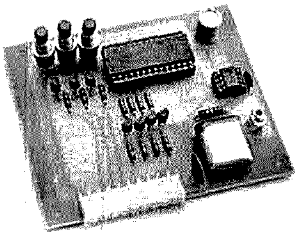
More Details? CHECK—OFF Page 126

outputs can be on somewhat different parts of the band with some reduction in gain. Standard models are available for any segment of the 26-230 MHz vhf range.

Price of the P13 Receive Multicoupler kit is $12.95. The P24 wired and tested model is $24.95, and the P50 model, which is housed in a die-cast box with BNC connectors, is $49.95.

For a complete catalog of vhf and uhf modules, send a self-addressed stamped envelope to Hamtronics, Inc., 182 Belmont Road, Rochester, New York 14612.

## real-time clock



Does your computer know what time it is? If not, you should add this new hardware peripheral from TED that will allow your computer to keep track of the time of day for control, timer and game applications. The *Real Time Clock* employs the latest cmos technology, resulting in high reliability, small size and low power consumption. The inputs and outputs are TTL-compatible and provide simple connection to your system with four wires from a parallel output port, ground, positive 5 volts, and positive 12 volts.

The *Real Time Clock* may be used with any computer system and includes complete hardware and software documentation and features crystal-controlled accuracy with a trimmer to allow exact frequency setting. Push buttons allow rapid and easy time setting. The clock can be operated from a separate power supply or battery, permitting it to keep the correct time, even when the rest of your computer system is off.

The crystal oscillator operates at a frequency of 3.579600 MHz and is divided by a factor of 59,600 to yield an exact 60-Hz frequency which is then applied to a digital clock chip containing the components to generate

a six-digit 12- or 24-hour time display. The time information from the clock is transferred in BCD form to the computer, one digit at a time. The digit-select inputs determine which of the six digits representing the time is present on the BCD output pins. A short program in machine language or BASIC may be used to transfer the time information from the *Real Time Clock* to the computer.

IC sockets are used for all of the ICs, and components are mounted on top-quality G-10 printed circuit board material. The unit is available in wired and tested form *only*, at a special introductory price of $39.95 plus $0.75 postage. Wisconsin residents please add 4% sales tax. For additional information, write TED, Box 4122, Madison, Wisconsin 53711.

## two-meter frequency synthesizer


VHF Engineering announces their new *SYN II Synthesizer*, a high-quality synthesizer designed for use in virtually all two-meter rigs available on the market today.

The SYN II Synthesizer is designed to operate over a frequency range of 140 to 149.995 MHz in 5 kHz steps and is compatible with two-meter equipment using transmit crystals in the 6, 8, 12, or 18 MHz range, and receive crystals in the 15 or 45 MHz range. The synthesizer may be used with either fm or phase-modulated transmitters and with mobile or base transceivers. The SYN II features unique i-f programming which permits the unit to be used with receivers having i-fs in the range of 100 kHz to 30 MHz. Detailed programming instructions are given in the construction manual so that the builder may select the standard 10.7 MHz i-f or any other i-f frequency in the permissible range, permitting use of the SYN II with older commercial equipment as well as currently available two-meter units. Standard repeater offsets of +600 kHz and −600 kHz are provided, along with

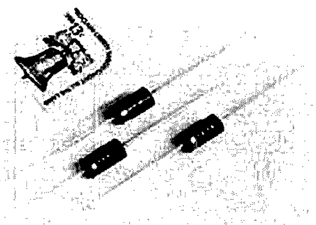


More Details? CHECK—OFF Page 126

three user-selectable offsets in 100-kHz steps, permitting the user to operate into standard repeaters or with unique offsets. A modification kit is available for MARS and CAP offsets.

The synthesizer kit consists of high-quality epoxy-glass circuit boards, computer grade components, thumb-wheel switches, stylized cabinet, and a detailed construction manual. The kit is complete and requires no additional components. The SYN II is available from dealers nationwide or direct. The price of the kit is $169.95; wired and tested $239.95; programmed to your equipment $249.95. For additional information, write VHF Engineering, 320 Water St., Binghamton, New York 13902.

## transient voltage suppressors



TRW has introduced a series of zener transient voltage suppressors (TVP) that protect circuits, systems, and equipment from voltage surges. The TVP1500 series rapidly changes the impedance value from a very high standby value to a very low conducting value when subjected to high energy transients. It shunts potentially damaging effects by clamping the voltage at some predetermined level.

The device features 1500 watts peak pulse power, a voltage breakdown range of 8.2 through 200 volts and a fast recovery time of approximately 1 picosecond. Series TVP1500, packaged in a polymer silicone case capable of withstanding temperature extremes to 400°C, also has a reverse standoff voltage range of 6.6 through 171 volts. All the devices in the series are 100 per cent surge tested, and can be used in telecommunication, automotive, computer, and dc power supply applications. The device is also available in 500 and 1000-watt peak pulse power ratings.

For more information write John Gamet, TRW Capacitors, 301 West "O" Street, Ogallala, Nebraska 69153.



More Details? CHECK—OFF Page 126

$1.50

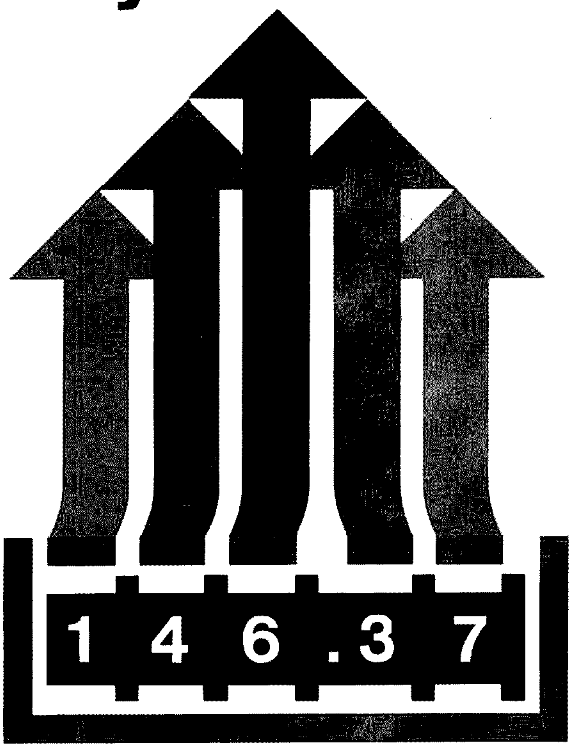
# ham radio

magazine

## AUGUST 1977



# direct output two-meter synthesizer



# ham radio magazine

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**volume 10, number 8**



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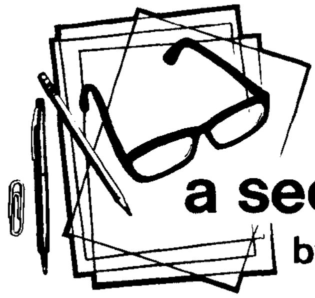


# a second look

by Jim Fisk

**As more and more amateurs** switch to factory-made gear, and as industry uses more ICs and disposable plug-in modules, the life of the dyed-in-the-wool ham homebrewer gets tougher and tougher. If you've recently tried any of the construction articles in the amateur magazines, you are already well acquainted with the hassle involved in obtaining a few needed components.

At one time you could drop in at your local corner radio store with a list of parts and the man behind the counter would fill your order. But that was when the vacuum tubes, resistors, and capacitors in your ham gear were the same as those in the family radio. It's not the same anymore — now the transistors and ICs in the radios and television sets are designed specifically for that purpose and have operating characteristics that are of little use elsewhere. There are exceptions, but they are few and far between.

Another problem that faces the serious home builder is the tremendous variety of transistors and ICs available from different manufacturers. Although some types of devices are made by more than one company, in most cases the semiconductor manufacturers crank out devices that are completely different from those of their competitors. And to add insult to injury, the same device may carry a dozen different part numbers: a 2N number, a replacement number, plus special numbers for units sold in large quantities to equipment manufacturers.

There is only one way to combat this lunacy: arm yourself with a good semiconductor cross-reference guide and a wide selection of electronic parts catalogs. Tops on the list of replacement guides is Howard Sams' *Transistor Substitution Handbook* available from *Ham Radio's* Communications Bookstore. This handy little paper back, which is updated every year, covers practically every transistor ever made, from 2N34 to 2N6500, with recommended substitutes. It also covers devices from Japan and Europe, as well as replacement types manufactured by Delco, General Electric, International Rectifier, Motorola, RCA, Semitronics, Sylvania, and Workman. Most of these manufacturers also publish replacement guides, available for the asking from their authorized distributors.

If you live in a large metropolitan area, chances are that there is an industrial electronics supply house that can fill your parts needs. Many of these firms don't advertise because they are not particularly interested in small quantity sales, but if you show up at their office, they will sell you the parts. If you want to find them, pick up your telephone directory and check the *Yellow Pages*: look under "Electronic Equipment and Supplies."

If you live out in the sticks, the problem is more difficult, unless you can get into the city. If you can't, you must purchase your components through the mail. Allied Electronics is the best bet in this case and you can get a catalog from any Radio Shack store. Be sure to get their industrial catalog though — the more common entertainment catalog is devoted primarily to CB and hi-fi and lists few electronic parts for amateur communications equipment.

**Jim Fisk, W1HR**
**editor-in-chief**

# presstop

IMMINENT COMMUNICATOR LICENSE action is being rumored by several industry sources. Departing Chairman Dick Wiley's support of the Communicator concept and his reported desire to see it realized before he steps down is one very strong agrument; its probable impact on the Personal Radio Division's budget, which will be reviewed shortly along with next year's proposed budget, is another.

It Appears Very Likely that the Communicator license will see some sort of official review within the next few weeks or so. What will come out of that review is another question.

EXISTING AMATEUR TRANSMITTERS WERE "GRANDFATHERED" June 2nd by an FCC modification to the first Report and Order on Docket 20777 that had become effective April 15. Under the modification all Amateur transmitters and transceivers (but not amplifiers) manufactured before April 15 are permanently exempted from the Report and Order's harmonic and spurious specifications. All Amateur equipment made after April 15 must meet the new specs, of course, but existing new equipment made before that date can be marketed until January 1, 1978. Individual Amateurs, however, are still responsible for meeting the 40-dB harmonic and spurious specifications of the FCC's first Report and Order on Docket 20777 in the operation of their own stations, even though the equipment itself has been grandfathered. The FCC's June 2nd relaxation applied only to the sale of non-complying equipment, and users are still expected to use it in such a way (with appropriate filters or an antenna tuner) that their stations meet the tighter requirement. Officially, the relaxation became effective July 18th.

UNRETURNED NOVICE EXAMS are still a big problem with the FCC in Gettysburg despite the dropping of multiple-exam mailings. Volunteer examiners have a major responsibility to see that a Novice exam, whether or not the applicant actually takes it, is returned to Gettysburg on time. Failure to do so can jeopardize the volunteer examiner's own license, and continuation of the present "unacceptable" number of unreturned exams could trigger drastic changes in Novice licensing!

GETTYSBURG RECEIVED A REPRIEVE when a radical personnel cut scheduled for June 10th didn't come off. Best news of all is that the previous "temporary" positions the people leaving had held are to be made permanent, and those people who have been filling the slots so well, will be staying in their jobs and working into permanent status.

The Reprieve Doesn't solve all of Gettysburg's problems, however. The Amateur workload continues to increase, and an estimated 10-20 additional people are going to be needed if the facility is to keep working smoothly.

"GUILTY ON TWO COUNTS" was the verdict the jury handed down June 6th in the trial of FCC Special Licensing Chief Richard Ziegler (July Presstop). One of the original four counts of bribery for the issuance of special Amateur callsigns was dropped and the jury failed to reach a decision on the second during the two-day trial.

18-YEAR-OLD GENERAL CLASS, or higher, Amateur license holders were permitted to administer Novice exams, effective June 13th. The amendment to Section 97.28(b) of the rules came about as a result of a Petition for Rule Making filed by WB4EKC.

A NEWLY-UPDATED EDITION of the FCC's Amateur Radio Rules, including all Part 97 changes through March 7, is now available from the Superintendent of Documents, U.S. Government Printing Office, Washington D.C. 20402. It's stock number 004-000-00338-1 and postpaid price is $1.30.

EXTENSIVE ELECTRONIC CONTROLS used in 1977 autos are causing RFI problems — a recent Illinois Bell notice warned that the "cruise control" in 1977 Cadillacs (and presumedly other GM cars) is sensitive to strong RF fields, which could cause sudden speed up or slow down. Some electronic skid control braking systems have locked up from RFI, and complete engine failure in fuel-injected engines has been reported by two-meter users.

A BOOKLET PROMOTING CANADIAN Amateur Radio has been published by the Radio Society of Ontario, Inc. The very attractive publication is available free to Canadian clubs or groups wishing to use it — an SASE to RSO, Box 334, Station U, Toronto, Ontario M8Z 5P7 will bring a sample and ordering information.

COMPUTER VOICE GENERATOR shown at the Dayton Hamvention is being used on 6 meters by WB4IVG in Dalton, Georgia. K1ZZ was "its" first contact.

# direct output two-meter synthesizer

New techniques permit the construction of a synthesized vhf transmitter which does not require frequency multiplication

**This article will describe** a unique, to amateur radio, method of building a two-meter synthesizer. Rather than function as a replacement for a crystal, *direct synthesis* generates the desired frequency without multiplication. Using ECL, TTL, and CMOS integrated circuits, the completed transmitter will produce 800 individual frequencies spaced every 5 kHz between 144 and 147.995 MHz. In addition, a local oscillator output, 10.7 MHz above the transmitter, can be used for receiver injection. With a phase-locked loop (PLL) ultimately controlling the vco (voltage-controlled-oscillator), the frequency accuracy is determined by a single crystal.

Contrary to some synthesizer designs, the receive and transmit frequencies in this unit are totally independent. This eliminates problems when odd frequency splits are encountered. Also, the two frequencies are available as BCD data for further processing or for a convenient readout rather than the thumbwheel switches. Again, another step into the realm of microprocessor-controlled equipment! The total cost for this 15-milliwatt exciter and local oscillator is approximately 100 dollars.

### frequency generation

The two-meter fm frequencies are all multiples of 5 kilohertz; therefore, with an accurately generated 5-kHz reference frequency, each channel can be produced through multiplication by using the proper integer (**fig.** 1). However, there are inherent problems in this scheme, primarily because no easily programmed frequency *multipliers* are available. On the other hand, programmable *dividers* do exist. By inserting the correct number of dividers (counters) into a feedback loop as shown in **fig. 2**, we have effectively created a frequency multiplier; this is the beginning of our PLL. Unfortunately, this method has several problems. For the output to be exactly on frequency, the difference detector must be driven to zero. Therefore, the detector must not have any offset; in addition, the error amplifier should have infinite gain. To overcome these problems, a PLL uses the phase of the vco as the controlling factor rather than its frequency. The phase detector will be discussed in more detail later.

There are two special problems in PLL frequency synthesis. Extreme care must be taken in the design to prevent the radiation of excessive sidebands and spurious outputs. The reference sidebands are caused by the vco being modulated at the sampling rate of the phase-detector. The more difficult (to control) spurious outputs are caused by close physical proximity to the digital logic. Also, the spurs can cause birdies in a companion receiver. Proper mechanical design, however, has reduced the levels to −55 dB and −90 dB, respectively.

### programmable divider and prescaler

Unfortunately, since we are working with a closed loop system, any problems in one area are reflected in other portions of the circuit. Consider the final output signal; there should be a minimum of buzz or hum associated with the signal. Also, the transient response (time to settle after a channel change) should be small. These problems can be reduced by

**By Jerry Pulice, WB2CPA**, 143 Gibson Avenue, Staten Island, New York 10308

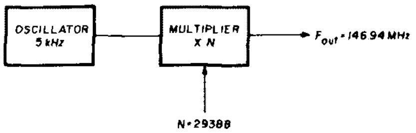



**fig. 1. Block diagram of a basic frequency synthesizer that uses integer multipliers to generate the output frequency.**

using a high performance loop to control the vco. As an example consider the PLL shown in **fig. 3**. With the vco prescaled by a factor of 20, the output frequency is determined by

$$F_{OUT} = 20(N)(0.005\ MHz) \tag{1}$$

Therefore, the minimum channel spacing will be 100 kHz instead of 5 kHz. To regain the original channel spacing, it would be necessary to divide the reference by a factor of 20. With a low reference frequency (250 Hz), the vco must have exceptional stability

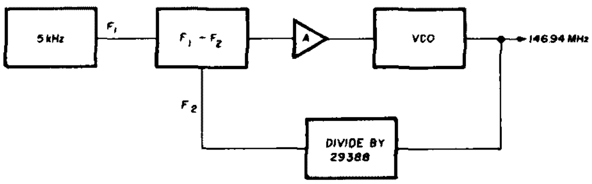



**fig. 2. This system eliminates the multipliers and uses more commonly available dividers to generate the required integer.**

since it can only be corrected 250 times per second. From these facts it can be seen that for best loop performance the digital logic should operate at the highest possible frequency.

TTL logic cannot operate directly at 144 MHz, and prescaling by at least 10 is required for any two-meter synthesizer. Conventional designs call for a programmable divider that can be preset to some number other than zero to modify the count length. Regardless of the counters toggle speed, this method will limit the upper frequency to approximately 15 MHz because not enough time elapses during one clock cycle to guarantee presetting the counters. It would appear that these problems would require the reference frequency to be lowered regardless of the loop considerations. However, return to **eq. 1**, which is one step beyond the basic form

$$F_{OUT} = (N)(M)(0.005\ MHz) \tag{2}$$

where $N$ = integer number
$M$ = prescale factor

Rearranging **eq. 2** yields

$$N = \frac{F_{OUT}}{(M)(0.005\ MHz)} \tag{3}$$

Where $N$ is now the number of divisions required by the prescaler and reference frequency. With a prescaler that divides by 20, $N$ would be 1440 divisions at 144 MHz.

Now, consider a prescaler that can divide by not only $M$, but $M+1$. To maintain the same output frequency, **eq. 2** must be rewritten to account for $M$ and $M+1$:

$$F_{OUT} = [M(N-A) + (M+1)(A)]\,(0.005\ MHz) \tag{4}$$

where $N$ = total number of divisions (integer number)
$M$ = prescale factor
$A$ = number of divisions at $M+1$

or $$144\ MHz = [20(1440-0) + 21(0)]\,(0.005\ MHz)$$

Reducing **eq. 4** to real components shows that a relatively slow counter (divider) can be used to control a fast two-mode (modulus) prescaler. In other words, the slow counter tells the prescaler to divide by twenty 1440 times and by 21 zero times. This technique is called pulse swallowing.

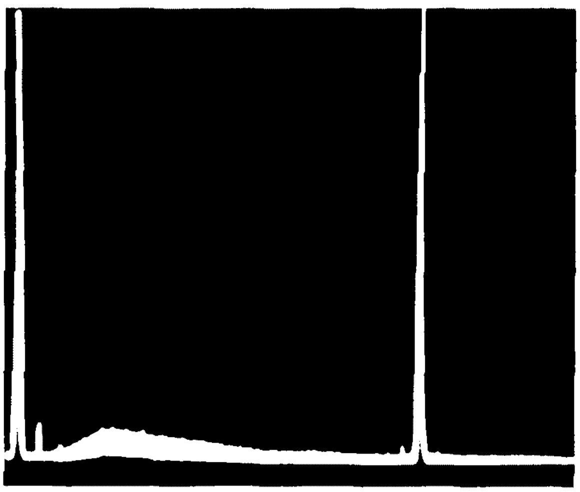
If one division though is by 21 instead of 20, **eq. 4** produces

$$[20(1440-1) + 21(1)]\,(0.005\ MHz) = 144.005\ MHz$$

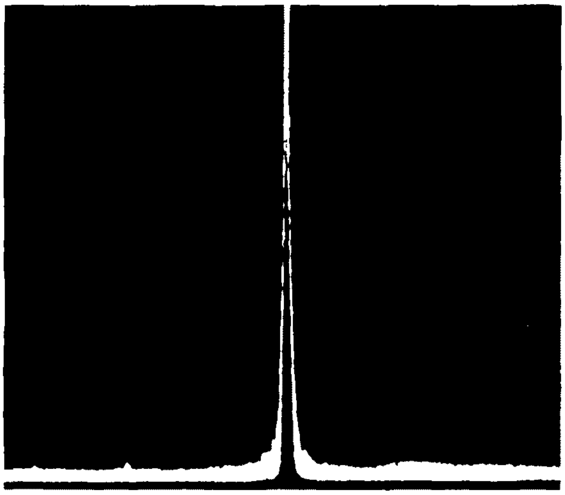
Therefore, for each division by 21, the vco frequency is raised 5 kHz. This relationship will continue each time the integer number is changed, producing channels that are separated by 5 kHz. The use of pulse-swallowing techniques overcomes the problems of vco and reference frequency, permitting the design of a vhf synthesizer with a channel spacing equal to the reference frequency.

## divider details

As shown in **fig. 4**, the programmable divider is



**As shown on this spectrum analyzer photograph, the output is very clean from 0 to 200 MHz. The signal frequency is 146 MHz. The low-frequency noise is being generated by the rf amplifier on the rf board.**



**Spectrum analyzer presentation of the final output. The signal is centered at 146 MHz. The left and right edges represent 143.5 and 148.5 MHz, respectively. The base line is approximately 80 dB down from the full output.**

split into a two modulus prescaler and a low-speed main counter. In conjunction with the 7474 flip-flop, the prescaler will divide by either 20 or 21, depending upon the level on the 95H90s SWALLOW ENABLE line. If this line is held high, one output will occur for each 21 input pulses. If the line is low, the output will be 1 pulse for each 20 input pulses. This output from the prescaler section then drives three synchronous-binary 4-bit counters arranged as a 12-bit binary divider. In transmit the counters are reset by U15 at $1479_2$ and in receive at $1586_2$ by U14. The initial input frequency is converted by the 7483s into data that presets the binary counters between 0 and 39. Assuming a transmit condition, the $1479_2$ count is shortened by the amount of preset, 0 to 39. The output of the counter will then occur at $1479_2$ to $1440_2$ representing 147.9XX to 144.0XX (**eq. 2**). It's now possible to generate any multiple of 100 kHz between the two frequency extremes.

Bit $1024_2$ (pin 12 U13) goes high once per count cycle and is used as the 5-kHz output to feed the phase detector. This pulse is much wider than the 5-kHz pulse at pin 9 of the binary counters and will provide a more reliable trigger for the CD4046 phase detector.

Unfortunately, the 100-kHz multiples have not fulfilled the requirement of 5-kHz channel spacing. Between 100-kHz increments there are nineteen desired channels spaced every 5 kHz. As determined by eq. 4, each time the prescaler divides by $M+1$ the vco frequency will increase by 5 kHz. Using 144.035 MHz as an example, the prescaler would have to divide by $M+1$ seven times and by $M$, 1433 times. To generate the required number of $M+1$ divisions, 10 and 5 kHz data, a rate multiplier is used. This device, comprised of a series of gates and counters, will produce a specific number of pulses on command. Controlled by frequency data (144.035 MHz) from thumbwheel switches, the rate multiplier will, each time it's enabled, generate seven pulses. The $1024_2$ bit from the binary counters is used to enable the multiplier. After 64 clock pulses EN OUT goes low, stopping the multiplier at $1088_2$. The rate pulses from U4 are temporarily held in a D-type flip-flop, U2. When released, these pulses are synced with the next clock pulse and also stretched into a full clock period for the SWALLOW ENABLE line. U2 is only enabled between counts $1024_2$ and $1088_2$, the same as the rate multiplier. The RS flip-flop (U3) controls U2. A timing diagram is shown in **fig. 5**. The use of hard-wired BCD data from the thumbwheel switches prevents the rate multiplier from generating more than 19 pulses for the prescaler.

## voltage controlled oscillator

One of the main criteria for vco design is that it be stable by itself. This synthesizer uses the Motorola MC1648 ECL logic oscillator (**fig. 6**). Of its many features, most important is its use of an external LC network as the frequency determining element. This type oscillator has less phase jitter than the RC switching oscillators (NE566 or MC4024).

The tank circuit consists of a Motorola MV109/209 varactor diode and a tuned line made from 3 inches (7.6mm) of miniature Teflon coax (RG-17U). This combination is extremely effective in combating the microphonics that plague other configurations. You

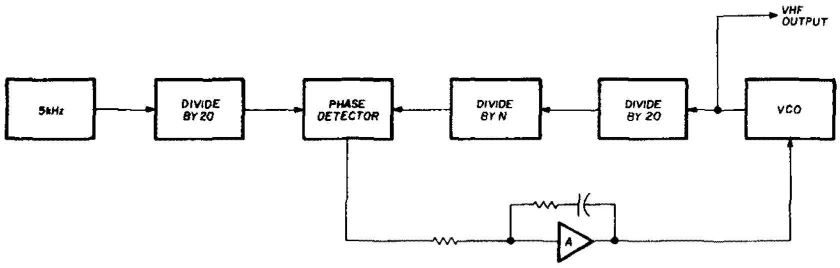
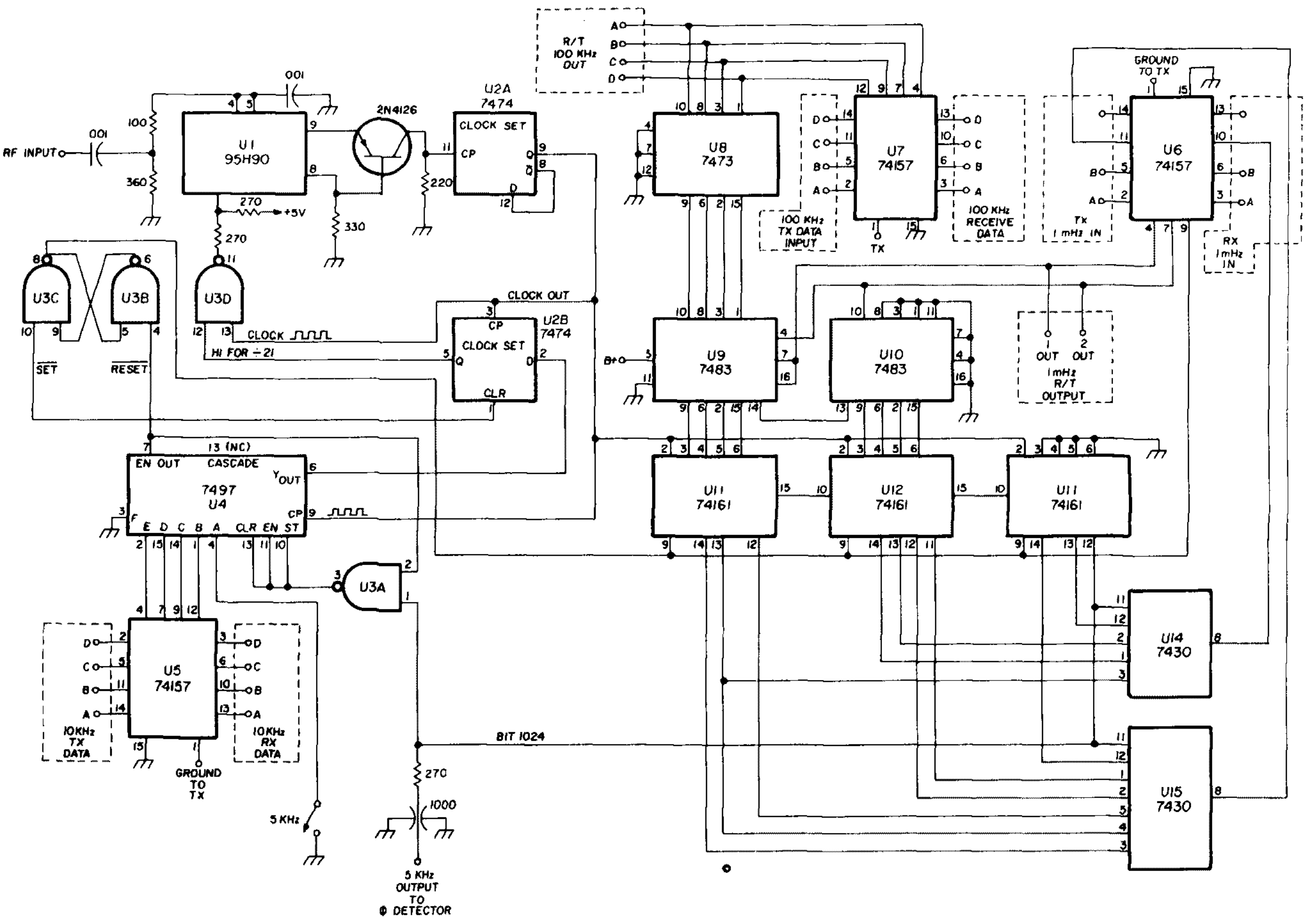
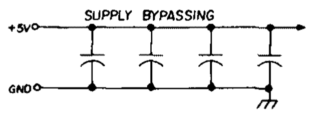



**fig. 3. Block diagram of the basic phase locked loop system. With a vco that runs greater than about 50 MHz, a prescaler is required for the loop.**




NOTE -

| IC NUMBER | GROUND | +5V |
|---|---|---|
| 7497, 74161/163, 74157 | PIN 8 | PIN 16 |
| 7400, 7430, 7474 | PIN 7 | PIN 14 |
| 7483 | PIN 10 | PIN 5 |

95H90 HAS POWER APPLIED

**fig. 4. Schematic diagram of the digital board. The prescaler section contains the 95H90 and the dual-D flip-flop. Pin 2, SWALLOW ENABLE, controls whether the prescaler is in the divide by 20 or 21 mode. The 0.001 μF capacitor connected to pin 5 of U1 should be mounted close to the IC. The $V_{CC}$ line requires bypassing with several values of capacitors to eliminate any switching transients from appearing on the line. Low-power Schottky ICs have been tested, and are recommended for replacement of the 74161/163, 7483, 74157, 7430, and 7474. When transmitting, the line marked TX should be grounded.**

can substitute any combination which will tune from 144 to 159 MHz with a tuning voltage of not less than 2.5 volts nor more than 10.5 volts.

The MC1648 drives three rf amplifiers in parallel. Each stage is untuned and delivers at least 20 milliwatts into a 50-ohm load. Even though hand-wound transmission line transformers are used, the lack of tuning makes the boards less prone to parasitics. One output drives the digital logic, another is used for the receiver local oscillator, and the last drives the transmitter. In this case, the "transmitter" is nothing more than additional power amplifiers.

A CMOS CD4046 is used as a frequency/phase detector. During a channel change, it departs from true phase lock and forces the vco to slew back to the correct frequency. When this point has been reached, the CD4046's output becomes a series of pulses with a duty cycle that is proportional to the phase difference. As soon as the phase difference also reaches zero, the output from the CD4046 enters a third state that effectively disconnects it from the loop amplifier. When correction is needed, the detector switches into the appropriate state.

The advantages of the CD4046 over a simple phase detector are many. Among these are the faster response to large channel changes and also lower

reference frequency sidebands in the rf output. A simple digital phase detector has a square-wave output with a 50 per cent duty cycle when locked. This represents no dc output so no further frequency change is required. However, there is a large ac output at the reference frequency which can be difficult to remove. In the design presented here, a locked condition is signalled by an open circuit from the detector which considerably simplifies the loop amplifier design. In practice the phase detector cannot operate at a zero phase difference and it has been set to produce an output pulse of approximately 5 per cent duty cycle.

The CD4046 drives a loop amplifier consisting of a MC1458V dual-operational amplifier. Despite what some articles on PLL would lead you to believe, more than just a lowpass filter is needed for optimum results. The first op amp could be classed as an integrator, but it also provides three time constants which insure stability of the loop. The second op amp is a simple 12 dB/octave lowpass filter which reduces the 5-kHz ripple on the tuning voltage. The simpler approach of using a lowpass filter versus a notch filter is justified since the performance improvement is very small.

The 5-kHz reference frequency is generated by the CMOS CD4060. This IC contains a 14-stage binary divider and three inverters for use as a crystal oscillator. The only other parts required are a parallel-resonant crystal and a few resistors and capacitors. The trimmer adjusts the crystal exactly to frequency and sets the final output accuracy.

## audio

To produce direct fm, the audio is summed with the tuning voltage after the op amps. The following requirements should be adhered to:

**1.** Audio compression and/or limiting should be used to hold constant deviation level.

**2.** Employ rapid rolloff past 3500 Hz to keep the radiated bandwidth narrow.

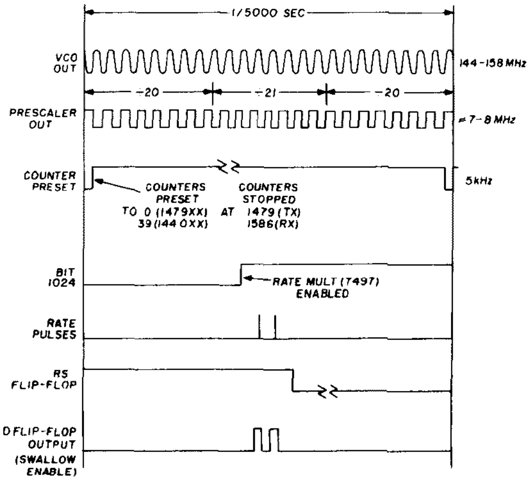
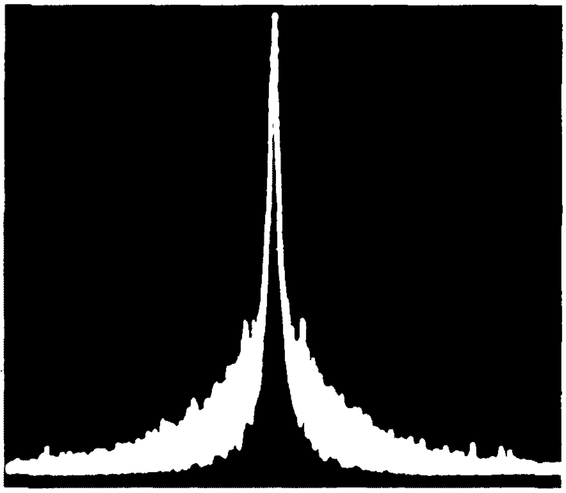



**fig. 5. Programmable divider timing diagram. This diagram represents a frequency of 14X.X10 MHz.**

**3.** Use standard 75 $\mu$sec pre-emphasis (**fig. 7**). This is direct fm and the audio will sound muffled if pre-emphasis is not included. The pre-emphasis should be applied after the clipping and filtering. There should be 1 volt p-p available after preconditioning the audio. As a warning, the vco has a sensitivity of 5 MHz/volt. Therefore, it only takes 1 mV of noise in the system to make this synthesizer useless in a narrowband fm system. Ground noise, loops, and proximity to other systems can also cause problems. The vco assembly *must* be housed in a completely sealed metal box. Diecast aluminum boxes are the best. Attempts at ultimate miniaturization will only produce 5 kHz whine, and possibly 60 Hz hum on the signal.



**This spectrum analyzer photograph shows the output within 50 kHz of the center frequency. The 5-kHz sidebands can be seen to be 54 dB down from full output.**

## circuit board checkout

**Digital card.** The initial testing of the digital board can be done at a frequency that will allow the pulses to be seen without relying on a high performance oscilloscope. An approximately 0.8 V p-p, 10-MHz signal should be injected into pin 1 of the 95H90. With power applied, the prescaler will run warm and total current drain will be about 600 mA for standard TTL. The complementary output pins, eight and nine, of the 95H90 should show an ECL level square-wave (3.3V to 4.1 V). The waveform will alternate be-

tween one tenth and one eleventh of the input frequency, which will cause blurring of the oscilloscope display. With the rest of the prescaler section working properly, the binary counters (U11-U13) will have a clock input on pin 2 that alternates between 1/20 and 1/21 of the input signal. If multiples of 100 kHz are selected, the input to the binary counters will be 1/20 of the input frequency; otherwise blurring will occur.

When it has been confirmed that the prescaler is functioning properly, the divide by $M+1$ function can be disabled by applying 5 volts to pin 3 of the 95H90. If gates U14 and U15 are operating correctly, a narrow negative-going pulse will appear on pin 9 of the binary counters. This pulse is used to load the binary counters after each cycle has been completed, $1479_2$ for transmit and $1586_2$ for receive. The ratio of the pulses at pins 2 and 9 of U11-U13 will correspond to the channel selected, 1440 to 1479. If the transmit line is high, 107 will be added to the ratios. This process can be extended to the entire programmable counter. With the 95H90 input as B and the pulses at pin 9 of the counters A, the ratios will be:

| | |
|---|---|
| 144.00 MHz Transmit | B/A - 28800 |
| 144.005 MHz Transmit | B/A - 28801 |
| 146.940 MHz Transmit | B/A - 29388 |
| 146.940 MHz Receive | B/A - 31528 |
| 147.995 MHz Receive | B/A - 31739 |

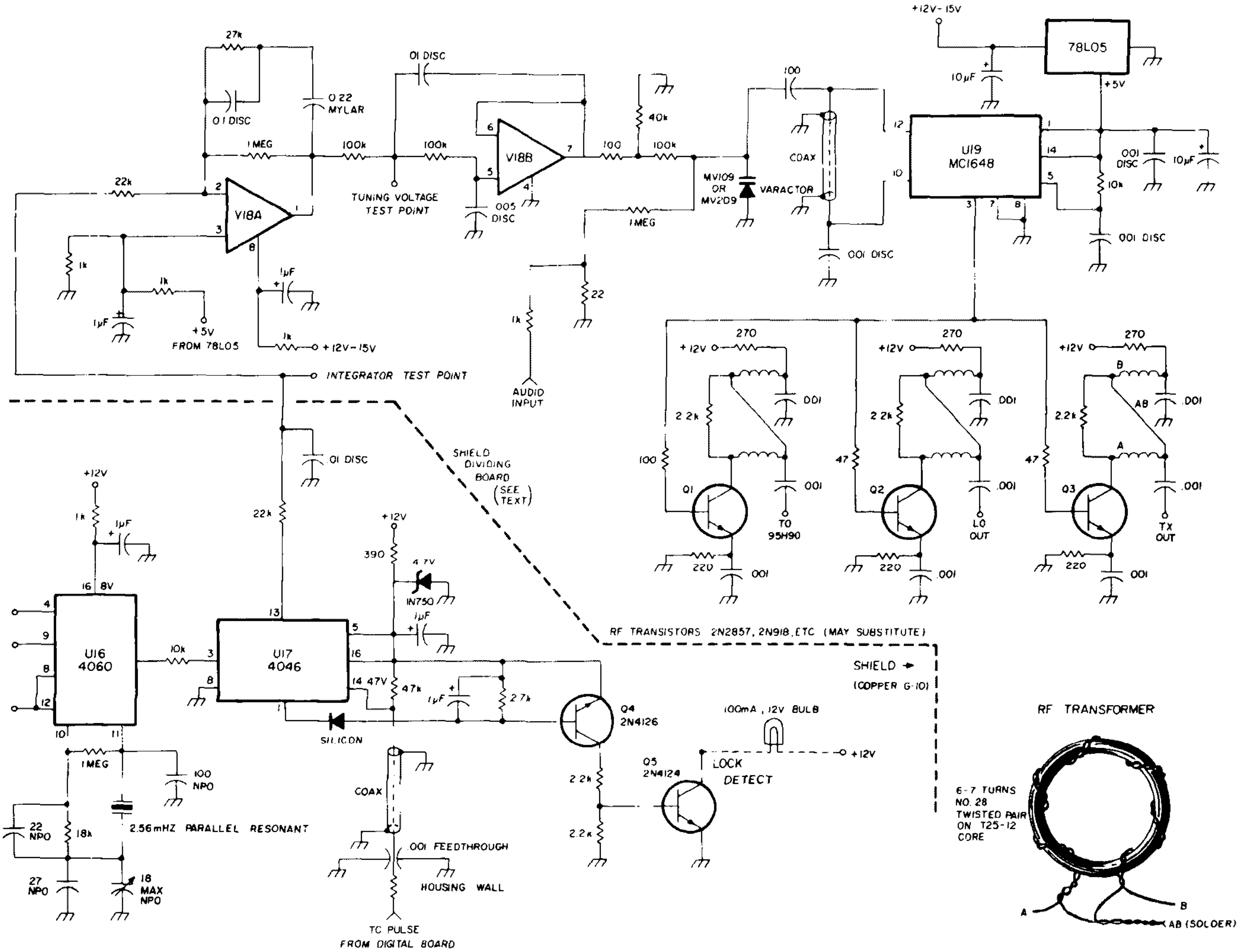



**fig. 6. Schematic diagram of the rf circuitry for the synthesizer. All resistors are ¼ watt; the electrolytics are dipped radial lead tantalums. The coax line can be replaced with 3 turns of number 22 AWG (0.6mm) wire. It should initially be wound to be ¼-inch (0.6mm) in diameter. The final configuration will be dictated by the required tuning voltage vs tuning range. The rf transformers are 6 to 7 turns of number 28 AWG (0.3mm) twisted pair wound on a T25-12 core. The 2.2k resistors across the transformers are soldered on the back of the printed circuit board.**

**Rf Board.** With power applied to the board from a 12-15 volt supply, the current drawn should be 60-75 mA. Check for the desired voltages on U16-U19 as shown on **fig.** 6. Testing of the vco and associated amplifiers will start at the vco and work back to the crystal oscillator. I recommend the Motorola MV109/209 varactor diode because of its wide $\Delta C$ range. This allows the vco to tune from 144 to 159 MHz with an input voltage of 2.5 to 10.5 volts. For test purposes only, the voltage can be supplied by a small adjustable supply connected to the test point just prior to the second op amp. Since the op amp has unity gain, correct operation of the vco can be determined from this point.

The integrator, first half of the MC1458, can be tested by grounding its inverting input. The vco should swing to at least 159 MHz when the junction of the two 22k resistors (integrator test point) is grounded. Conversely, it should move to below 144 MHz when 5 volts is applied to the same point.

At pin 4 of U16 there should be a 40-kHz square wave that can be used to set the crystal on frequency. By using the 125th harmonic of the 40-kHz signal, the crystal can be zero beated against the 5-MHz WWV frequency standard. The phase detector requires a 5-kHz input that is TTL compatible.

To test the phase detector, connect an NE555, or similar, oscillator to the 5-kHz pulse input on the rf board. With a pulse frequency less than 5 kHz, the vco should be driven to its lower frequency limit. If the pulse frequency is decreased below 5 kHz, the vco should swing to its upper limit. At this point the boards can be connected together forming an almost complete synthesizer.

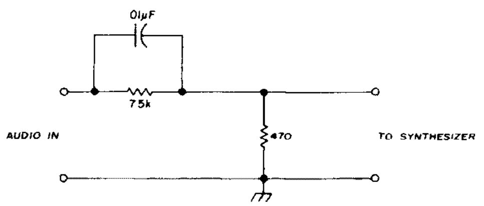



**fig. 7. Schematic diagram of a standard 75 μsec pre-emphasis network.**

**Switches.** Due to the method of generating the correct presets for the binary counters and inputs to the rate multiplier, two forms of input data were used, BCD and $\overline{\text{BCD}}$. The information for the 10-kHz frequencies is in the normal BCD form, while that for the 100 kHz and MHz is $\overline{\text{BCD}}$. Using the basic premise that an open on a line is equivalent to a digital 1 (high) the appropriate switches can be selected. It should be remembered that a $\overline{\text{BCD}}$ switch, not a BCD switch, produces four open circuits for a BCD zero. Regardless of the switch type you select, to use the switch in its true form connect the common terminal to ground and pull the four outputs to 5 volts through 4 to 10-kilohm resistors. To complement the switch, connect the common lead to 5 volts and pull each output to ground through 270-ohm resistors. **Table 1** shows the correct BCD input information. To add 5 kHz to the output frequency, the 5-kHz line should be taken high (1).

The complete synthesizer can now be tested. The appropriate inputs and outputs on the two boards can be connected with lengths of either miniature coax or twisted-pair cable. A small 12-volt lamp can be attached to the LOCK DETECT terminals. Since this is only a test, the output will not be very clean and the signal should not be put on the air. Connect one rf output to a frequency counter and terminate the third in 50 ohms. When power is applied to both boards the lamp should flash once and the synthesizer *should* be on the correct frequency. If the synthesizer does not lock at all, the lamp will remain lit.

**table 1. Synthesizer truth table**

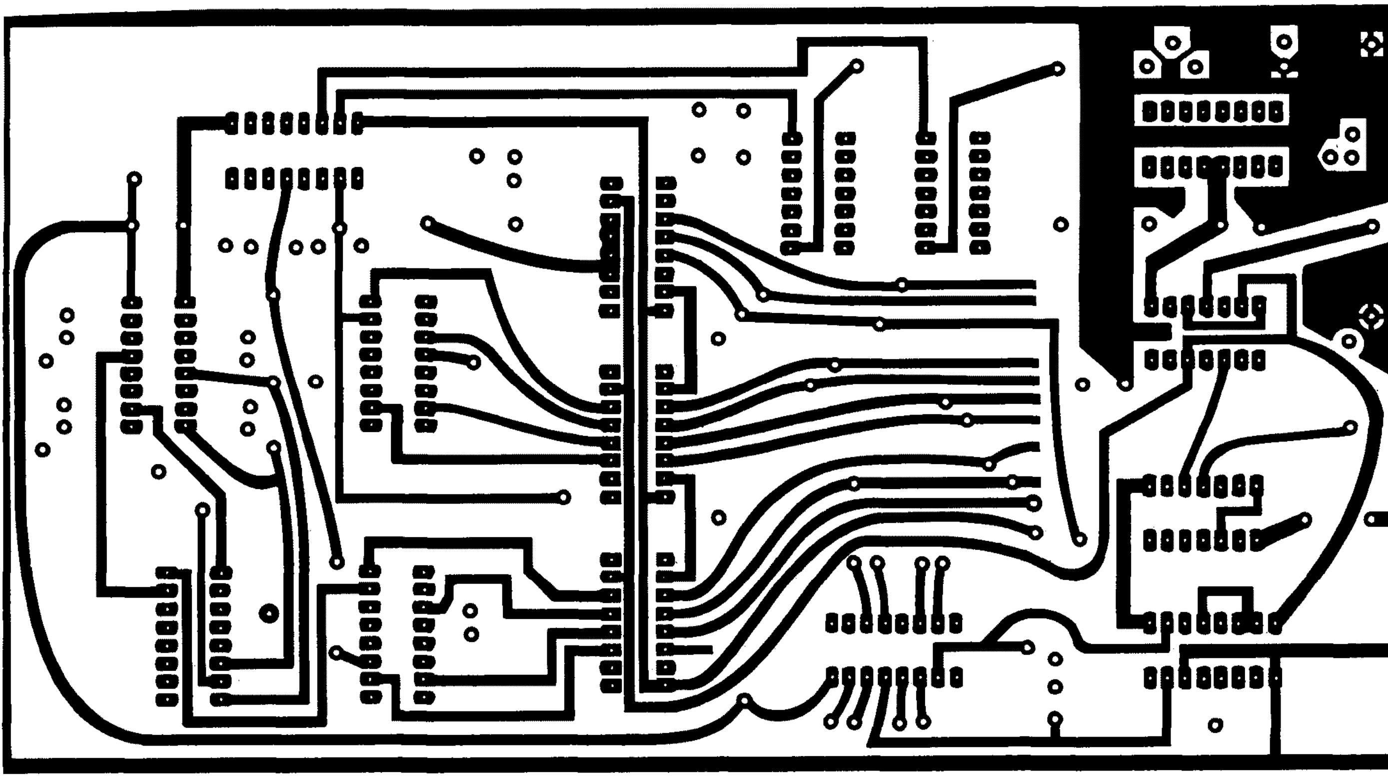
| 1 MHz | B A | 100 kHz | D C B A | 10 kHz | D C B A |
|---|---|---|---|---|---|
| 144 | 0 0 | 0.0 | 1 1 1 1 | 0.0 | 0 0 0 0 |
| 145 | 0 1 | 0.1 | 1 1 1 0 | 0.1 | 0 0 0 1 |
| 146 | 1 0 | 0.2 | 1 1 0 1 | 0.2 | 0 0 1 0 |
| 147 | 1 1 | 0.3 | 1 1 0 0 | 0.3 | 0 0 1 1 |
| | | 0.4 | 1 0 1 1 | 0.4 | 0 1 0 0 |
| | | 0.5 | 1 0 1 0 | 0.5 | 0 1 0 1 |
| | | 0.6 | 1 0 0 1 | 0.6 | 0 1 1 0 |
| | | 0.7 | 1 0 0 0 | 0.7 | 0 1 1 1 |
| | | 0.8 | 0 1 1 1 | 0.8 | 1 0 0 0 |
| | | 0.9 | 0 1 1 0 | 0.9 | 1 0 0 1 |

## troubleshooting

If there are problems, a return to the circuit board checkout phase might be appropriate. Remember that this is a feedback system and trouble in one section can cause apparent difficulty in another. A good way to troubleshoot the unit is to clamp the tuning voltage at some fixed value from an external power supply. If the vco will not tune within the desired range, lock cannot be achieved.

A more subtle problem is a locked synthesizer but with the output on the wrong frequency. The cause of this problem will be found on the digital board, assuming the initial 5-kHz signal is correct. Make sure that the components are soldered on each side of the printed-circuit board. Unless plated-through holes are used, it may be difficult to solder sockets on both sides of the board. One solution is to use Molex pins, another is to mount the sockets over spacers. If the synthesizer is still off frequency, observe the pattern of the errors.

**1.** No channel spacing less than 100 kHz. This means that the divide by 21 function of the prescaler is not being enabled. Start at pin 2 of the 95H90 and work back to pin 6 of the rate multiplier (U4).



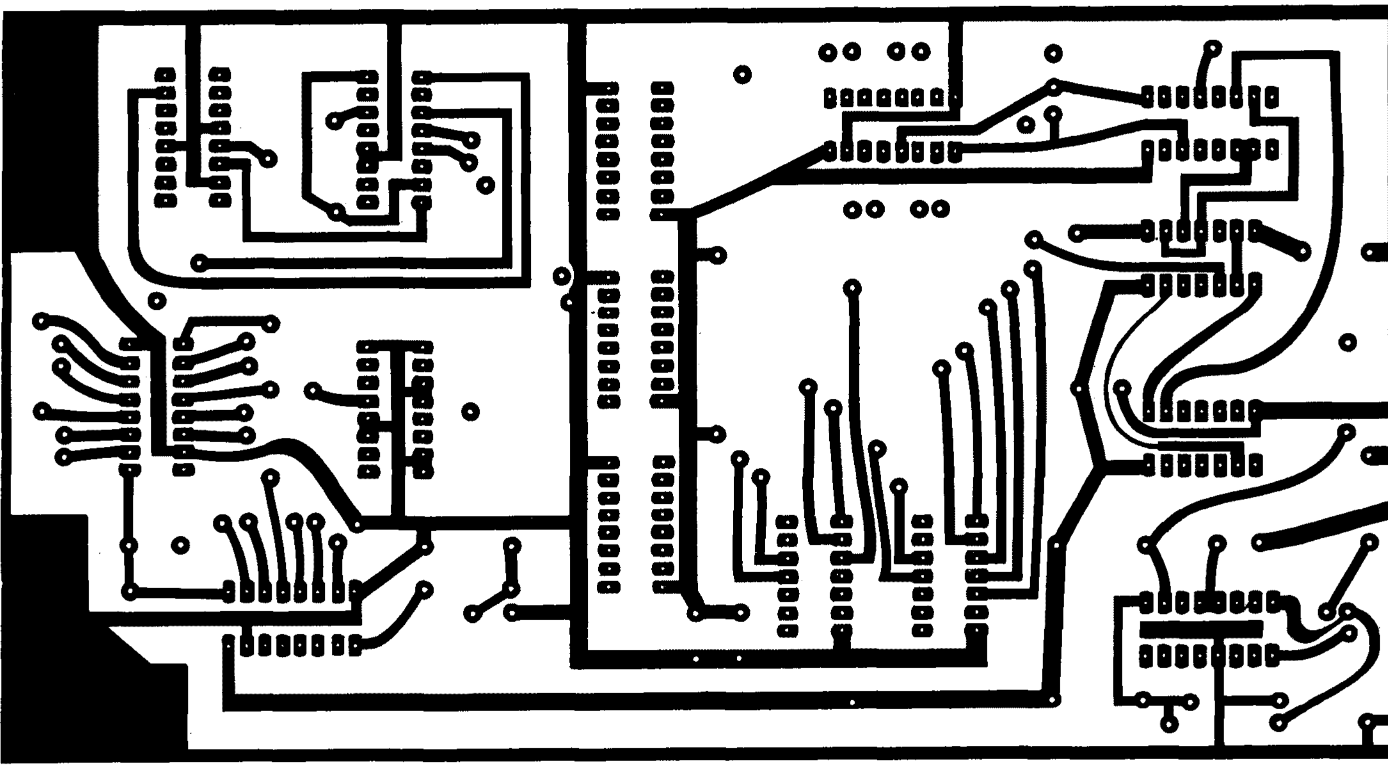


fig. 8. Circuit-board layout for the digital board. The top diagram shows the foil pattern for the top of the board, while the other side of the board is shown at the bottom.

2. Constantly high frequencies. The divide by 21 feature is being enabled too long.

3. Constant error in hundreds of kHz could mean that a 7483 full adder is faulty. They convert the $\overline{\text{BCD}}$ data into a binary code.

4. An output that is correct only on alternating binary increments means that a bit has been dropped between the 7483s and the preset inputs of the counters. For example, 144.000 to 144.200 is correct, 144.200 to 144.400 is wrong, and 144.400 to 144.600 is correct, etc.

5. If one of the 7430s (U14 or U15) is bad, either *all* the transmit or *all* the receive channels will be off.

Substitution is the best method of checking the ICs. You can save enormous amounts of time by mounting them in sockets, except the MC1648 and 95H90. If you don't have adequate test equipment, use a signal generator at a lower frequency to test the digital board.

## synthesizer related problems

With the techniques used to generate this type of synthesized equipment, there can be many problems that are system oriented. Some areas may require a look at the overall performance before the basic problem can be solved.

1. Sidebands at the TTL clock rate are caused by insufficient isolation between the circuit boards. If these sidebands are radiated by your transmitter, they are illegal; they will appear at the output frequency divided by 20, 144.00 ± 7.2 MHz, ± 14.4 MHz ± 21.6 MHz, etc. The general cure for this type of problem is to put the rf board into a sealed metal box (leads that enter the box should go through 0.001 $\mu$F feedthrough capacitors).

2. This synthesizer delivers 10 kHz p-p deviation with less than 0.2 volt input so it will be prone to overmodulation. The deviation can be set quite easily if you make use of the synthesized local-oscillator output. With this output connected to a receiver, modulate the synthesizer until the audio level is the same as a local repeater. It doesn't matter to the discriminator whether its's the actual incoming signal that is being modulated or the local oscillator. For normal operation though, use a relay contact to short out the audio line during receive periods. Otherwise, audio feedback will occur.

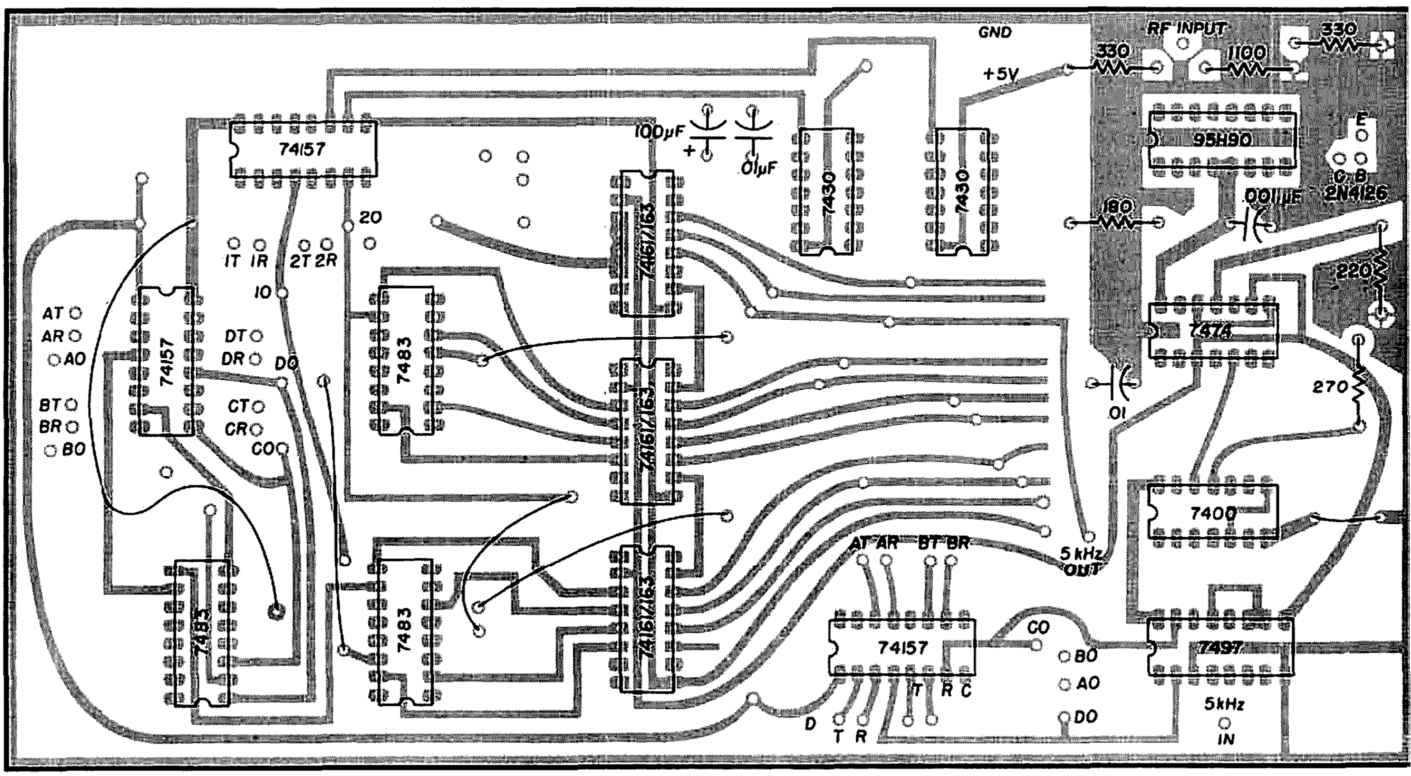



fig. 9. Component placement for the digital board.

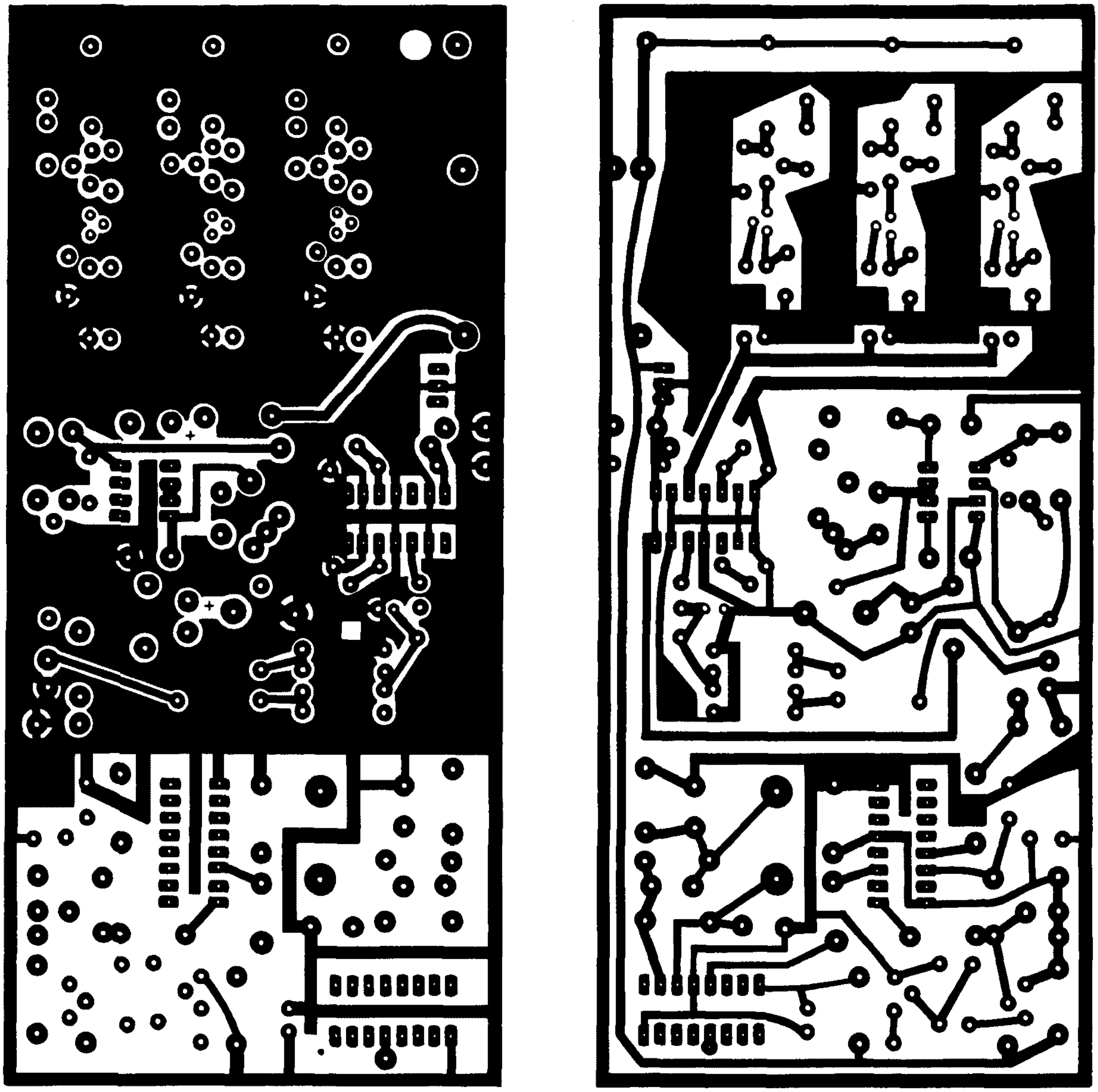


fig. 10. Circuit-board layout for the rf board. The top of the board is shown at the left and the bottom at the right.

**3.** Sidebands at 5 kHz cause annoying whine in your receiver and will be transmitted as well. Any sidebands experienced are caused by the physical placement of the rf board. Also, the shield between the portions of the board must be in place. This can be made from a piece of double-sided printed-circuit board. The output of the digital board must enter the enclosure through the resistor and capacitor combinatigns shown on the schematic diagram (**fig. 4**).

**4.** Receiver birdies are possible if the programmable divider is not shielded. Microphonics are not a problem and the enclosure does not have to be some type of sealed box. To avoid the problem of many feedthrough capacitors for the digital switches, I suggest that you mount the switches inside this box.

If you follow the previous suggestions, a clean transmitter should be no problem. I use four amplifier stages to directly increase the output to 40 watts. The LOCK DETECT output should be used to prevent keying the transmitter if the synthesizer is running wild.

## receiver interfacing

To be entirely free of birdies, on-channel spurious radiations must be on the order of 100 dB below the local-oscillator level. Achieving this requires constant attention to smalll details such as shielding, removal of ground loops, and maintaining isolation. Certain receiver designs can give you perhaps 20 dB margin against such birdies (this assumes the birdies are

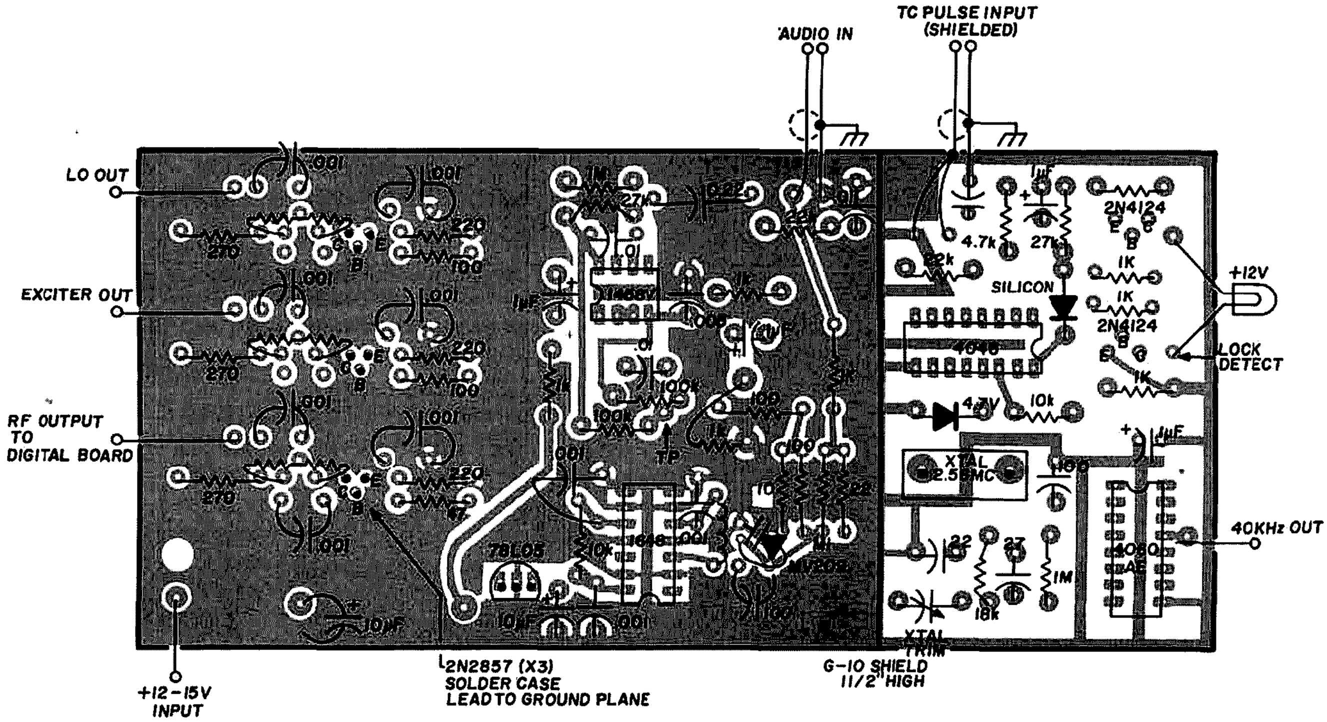



fig. 11. Parts placement diagram for the rf board. The resistor across the toroids is mounted on the rear of the board.

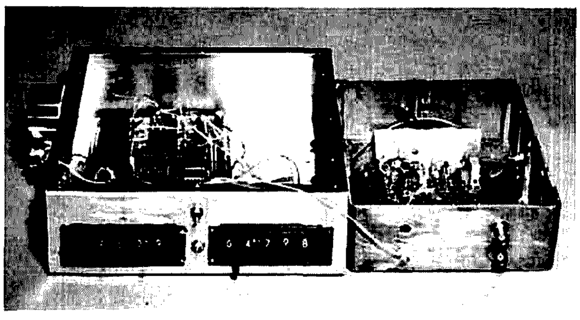
produced by the local oscillator and not radiated into the frontend).

A product or balanced mixer has the ability to reject certain forms of noise near the local-oscillator frequency. If you are contemplating the design of a companion receiver, think twice before going with the now standard dual-gate MOSFET mixer. A better choice might be a JFET mixer using a device such as the high $I_{dss}$ U310. A single-balanced mixer with this transistor will have a wider dynamic range as well as significant local-oscillator noise rejection. These items are important when operating in urban areas, or when trying to achieve 0.2 $\mu$V sensitivity on all channels.

No matter what first mixer configurations you select, two old design rules still apply.

1. Set your mixer so that conversion gain does not increase if you raise the local-oscillator power. If it does, the local-oscillator port is not saturated and the receiver will be overly sensitive to birdies and noise from the synthesizer.

2. The input ports of the first mixer should see only the minimum necessary bandwidth. If you don't filter the i-f output, you'll have no selectivity, obviously. Image and intermod problems will occur if the rf input isn't filtered. This includes both sides of the rf amplifier. If the local oscillator isn't filtered, you will have a severely degraded mixer noise figure. This applies more to single-ended mixers. Such a mixer acts as a high-gain amplifier to i-f frequency signals injected into either input. The solution is a highpass filter or, preferrably, a tuned circuit with a moderate loaded *Q* in the injection line. The tuned circuit



Photograph of the completed synthesizer. The rf circuitry is housed in the cast aluminum box on the right. The digital portion of the synthesizer is on the left.

serves the dual purpose of voltage transformation as well which can be useful in gate-driven fet mixers.

The previous comments should enable you to build a transceiver equal in all respects, except one, to a crystal-controlled unit. The final problem area is ultimate receiver selectivity. The noise sidebands of

the MC1648 do not fall off quite as fast as those of a well designed discrete vco. The difference is slight and transmitter performance is uneffected. Residual audio fm measures less than 100 cps in this design. However, adjacent channel rejection in the receiver will be degraded. In normal operation, crystal-controlled equipment would be able to maintain DX communications within 10 kHz of a repeater channel. This design will require a spacing of about 15 kHz from the active channel. The difference is not noticeable unless your receiver has a good quality i-f filter and the shielding necessary to make full use of it. Also, you'd have to check closely to notice the difference. It is worth mentioning that Heathkit chose the same vco for their down-converting multiple-crystal synthesizer; they were able to attain a 30-kHz selectivity of 60 dB minimum.

## convenience options

Because the synthesizer has the BCD frequency data available at all times, some other type of display can be connected. It would be much easier to see a seven-segment display in the dark than the thumb-wheel switches. In fact, you do not need switches at all. The channel information could be generated by a circuit that would scan a frequency range until a signal is detected. The combinations are almost endless. Remember though, both BCD and $\overline{BCD}$ data is required.

## conclusions

I hope that this article has removed some of the air of black magic that seems to have been associated with frequency synthesizers in the past. Unlike other articles, this one was concerned with how to get the radio to work, what you can expect to go wrong with it, and what you can do about it. Most of the literature on synthesizers has been an endless tirade about Laplace transforms and loop stability. These are important, but a critically damped response and the associated theoretical model by themselves only make a good BSEE senior term paper. This unit has been on the air for a year at WB2CPA, and successful duplication should not pose a problem to a reasonably competent amateur.

**ham radio**

# how to design Yagi antennas

Discussion of a new Yagi design method, developed at the National Bureau of Standards, which allows you to design Yagis for your own operating requirements with optimized, reproducible gain characteristics

**Have you ever wondered how to** design a really good Yagi for your own requirements rather than just guessing, or using an existing design? If so, this article should be just what you are looking for. By using the information presented here you can design your own optimum Yagi for any frequency, from hf through uhf, with booms up to 4.2 wavelengths long.

Up until now, there has been little design information for Yagi antennas in the amateur literature. Kmosko and Johnson[1] designed a 13-element Yagi at 144 MHz but gave information only on that specific model. Greenblum[2] provided ranges of design values but was not specific as to exact sizes. The tables from Greenblum's article have appeared in recent *ARRL VHF Handbooks* and *Antenna Handbooks,* and several amateurs have reported good correlation using the mean values specified. Recent articles in the professional journals (such as the *IEEE Professional Group on Antennas and Propagation,* and others) have published computer-aided designs, but specific *cook-book* information is not available.

Now, for the first time, a straightforward approach to Yagi designs of various sizes and gains is available.[3] It is the result of an exhaustive study by the National Bureau of Standards in the early 1950s to explore all the major antenna types (Yagis, corner reflectors, rhombics, etc.) suitable for use on vhf

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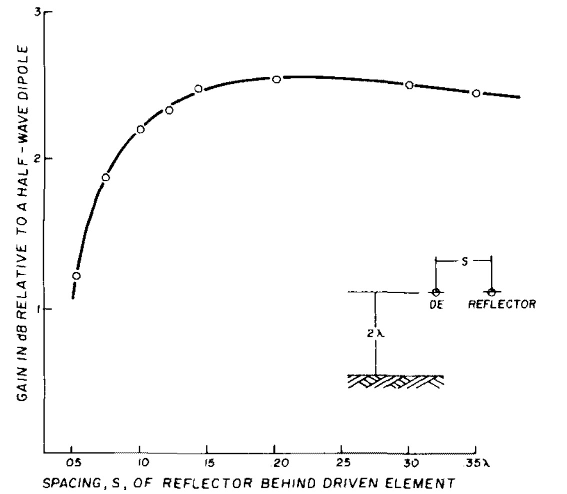


**fig. 1. Gain in dB of a driven element and reflector for different spacings between elements.**

ionospheric scatter. The NBS report tabulates all the design information necessary to construct six different boomlength Yagis (this portion of the project took nine man years to complete). The only known amateur use of these data are the WØEYE 432-MHz Yagi[4] and several unpublished Yagi designs by WØPW (ex WØEYE) and W1JR.

This NBS report shows the interrelationship between director and reflector diameters, lengths, and spacings, as well as the effects of a metal supporting boom. Optimum designs and gains for various boomlengths from 0.4 to 4.2 wavelengths are shown along with nomographs for designing a Yagi for your own operating requirements. Those readers who are interested in all the specifics will find the NBS publication invaluable. This article will highlight the results and present all the information necessary to design such Yagis; several working design examples will also be discussed.

## reflectors

During the NBS investigation into optimum Yagi design, various reflector lengths and spacings were tried on a two-element Yagi. As can be seen from **fig. 1**, maximum gain is 2.6 dBd, peaking broadly at 0.2λ behind the driven element. Hence, all the Yagi designs presented here are optimized using this reflector spacing.

The NBS engineers tried various other reflector configurations in order to realize any possible increase in gain. The trigonal configuration shown in **fig. 2** yielded the maximum increase, 0.75 dB over a single reflector, when tested on a Yagi 4.2λ long. It should be applicable to the other designs and may be desirable if high front-to-back ratios are desired.

The heart of any Yagi design is the director. Extensive tests have shown that the diameter, length, and spacings are all interrelated. Also, it should be pointed out that these parameters become increasingly critical as the number of directors (and hence the boomlength) increase.

NBS tested various director lengths using spacings of 0.01 to 0.40λ on booms to 10λ long. Plots of these combinations show that there are optimum spacings for maximum gain. As the boomlength is increased, the optimum director spacing also increases. In addition, the gain of the antenna can be further increased if the length of each director is carefully chosen. It is noted that the diameter of the element affects its length, thicker directors being shorter than thinner ones. A comparison of maximum gain versus boomlength for uniform and optimized length directors is shown in **fig. 3**. Those readers desiring further information are referred to *NBS Technical Note 688*.[3]

A set of optimum director and reflector lengths normalized to 0.0085λ diameter elements is

**table 1. Optimized lengths of parasitic elements for Yagi antennas of six different lengths (reflector spaced 0.2λ behind driven element, element diameter 0.0085λ).**

| | Length of Yagi in Wavelengths | | | | | |
|---|---|---|---|---|---|---|
| | 0.4 | 0.8 | 1.20 | 2.2 | 3.2 | 4.2 |
| **Length of Reflector, λ** | 0.482 | 0.482 | 0.482 | 0.482 | 0.482 | 0.475 |
| 1st | 0.442 | 0.428 | 0.428 | 0.432 | 0.428 | 0.424 |
| 2nd | | 0.424 | 0.420 | 0.415 | 0.420 | 0.424 |
| 3rd | | 0.428 | 0.420 | 0.407 | 0.407 | 0.420 |
| 4th | | | 0.428 | 0.398 | 0.398 | 0.407 |
| 5th | | | | 0.390 | 0.394 | 0.403 |
| 6th | | | | 0.390 | 0.390 | 0.398 |
| 7th | | | | 0.390 | 0.386 | 0.394 |
| 8th | | | | 0.390 | 0.386 | 0.390 |
| 9th | | | | 0.398 | 0.386 | 0.390 |
| 10th | | | | 0.407 | 0.386 | 0.390 |
| 11th | | | | | 0.386 | 0.390 |
| 12th | | | | | 0.386 | 0.390 |
| 13th | | | | | 0.386 | 0.390 |
| 14th | | | | | 0.386 | |
| 15th | | | | | 0.386 | |
| **Spacing between directors, in λ** | 0.20 | 0.20 | 0.25 | 0.20 | 0.20 | 0.308 |
| **Gain relative to half-wave dipole, dB** | 7.1 | 9.2 | 10.2 | 12.25 | 13.4 | 14.2 |
| **Design curve (see fig. 4)** | (A) | (C) | (C) | (B) | (C) | (D) |

presented in **table 1**. These data, with respective gains noted, yield optimum performance for the six boomlengths which are shown. If a different element diameter is desired (isn't that always the case?), the elements can be scaled by using the nomograph in **fig. 4**. Element diameters from 0.001 to 0.04λ can be easily scaled as will be discussed later.

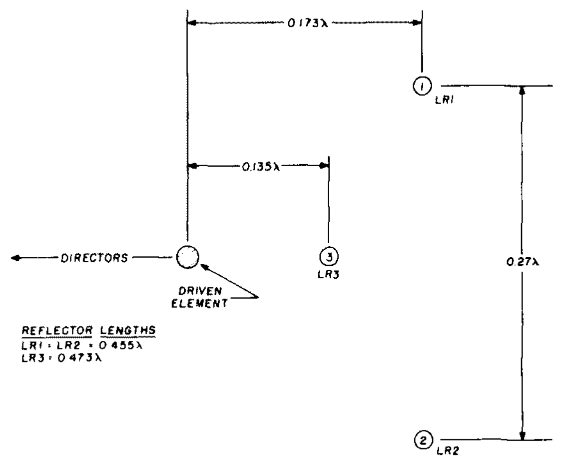


**fig. 2. Trigonal reflector arrangement (three reflector elements), when used with 4.2λ Yagi, provides 0.75 dB increase in gain (lengths not corrected for boom thickness).**

The element data presented in the NBS report is based on an *air* boom which, in the original tests, was simulated by a triangular plexiglass structure. After optimization was completed, various booms and materials were tested to check the effects of the test boom. All measurements verified that the designs tested on plexiglass were optimum in an air dielectric. However, attempts to repeat these results using wooden booms were dismal. According to Peter Viezbickie, the author of the NBS report, changes in moisture and directivity due to the wooden booms made repeatability almost impossible despite various coatings applied to the wood.

Metal-boom Yagis were entirely repeatable if the elements were lengthened to compensate for the boom structure. At first glance, it may seem that a constant factor could apply. However, tests conducted by NBS showed that small diameter booms (with respect to wavelength) had less effect on element lengths than larger booms. These data are plotted on **fig. 5** for boom diameters up to 0.04λ. Tests also showed that, for correction purposes, the effect of square and round booms were identical.

## feed systems

Detailed feed systems are not discussed in the report. On most tests, a folded dipole using a 4:1 half-wavelength coaxial balun was followed by a stub tuner. However, any of the usual feed systems can be used.[5] Reference 6 describes how to test these matching systems.

## patterns

Finally, the NBS report shows radiation patterns for the *E* and *H* planes. For the sake of brevity, only the patterns for the 1.2λ and 4.2λ Yagis are presented in this article (see **figs. 6** and **7**). You will note the symmetrical pattern, the low side lobes, and the high front-to-back ratio, all characteristics of a well-designed Yagi antenna.

Tests made by W6FZJ and W0EYE on a 15-element, 4.2λ Yagi for 432 MHz, designed with the method described in this article, showed that the antenna had about 1% vswr and 1-dB gain bandwidth, slewed to the low-frequency side of the center design frequency; performance above the center frequency fell off quite rapidly. It is estimated that the gain and vswr bandwidth for the 1.2λ Yagi is about 2%. It should be pointed out that the bandwidth of a Yagi is quite often limited by the matching and feed system, not by the basic Yagi design. In this respect most amateur beams use narrowband feed systems compared with Yagis designed for use in commercial service.

## designing a yagi antenna

We will now proceed to design a 1.2λ Yagi for 50.1 MHz, a 2.2λ Yagi for 205 MHz, and a 4.2λ Yagi for

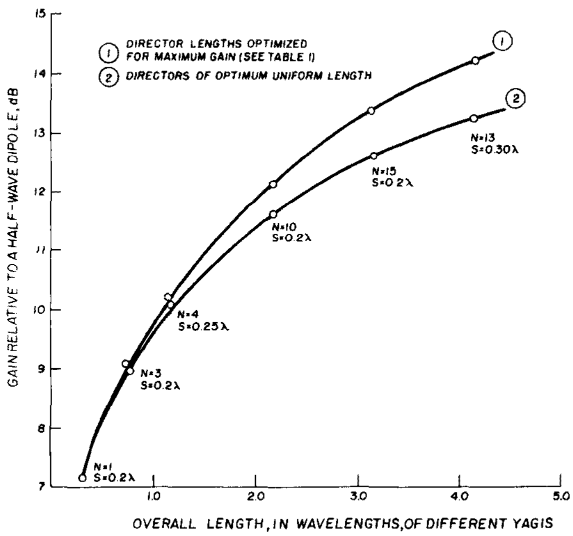


**fig. 3. Gain comparison of different length Yagis, showing the relationship between directors optimized in length to yield maximum gain, and directors of optimum uniform length. *N* is the number of directors; *S* is the spacing between directors (reflector spaced 0.2λ on all antennas).**

432 MHz to demonstrate how the NBS design material can be used. I actually built and tested each of these designs to verify the validity of the design data. In all cases the performance of the finished antennas matched the results reported by NBS.

The first step in any design is to choose the desired gain, compare it with the designs in **table 1**, and see if the stated boomlength is within the desired range. Next, the element diameter should be chosen to fall within the specified ranges (0.001 to 0.04λ) on the design nomograph, **fig. 4**. Finally, the boom or supporting structure should be chosen.

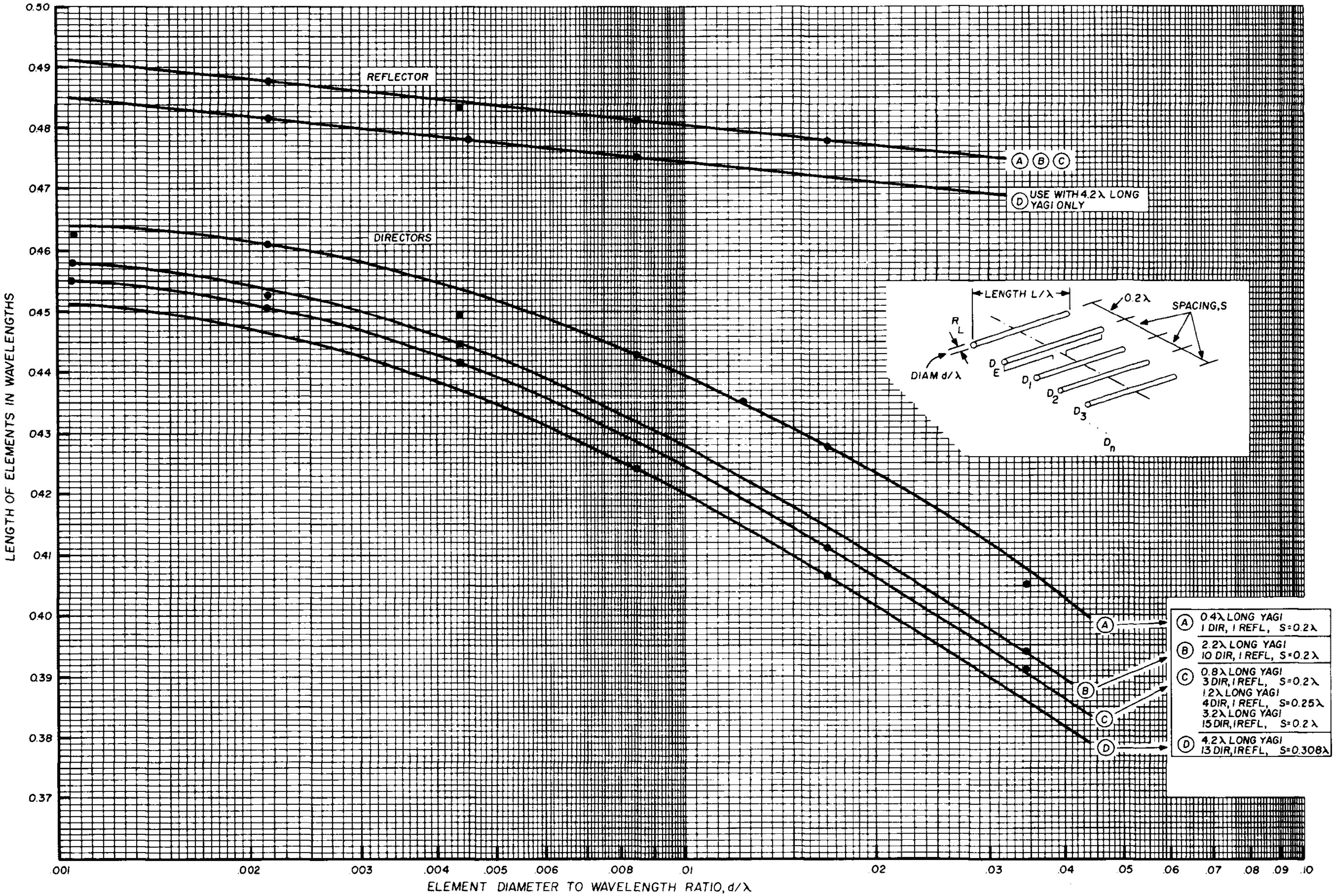


**fig. 4.** Yagi antenna design nomograph showing the relationship between element diameter to wavelength ratio (d/λ), and element length for different antennas. Detailed procedure for using this chart is presented in the text.

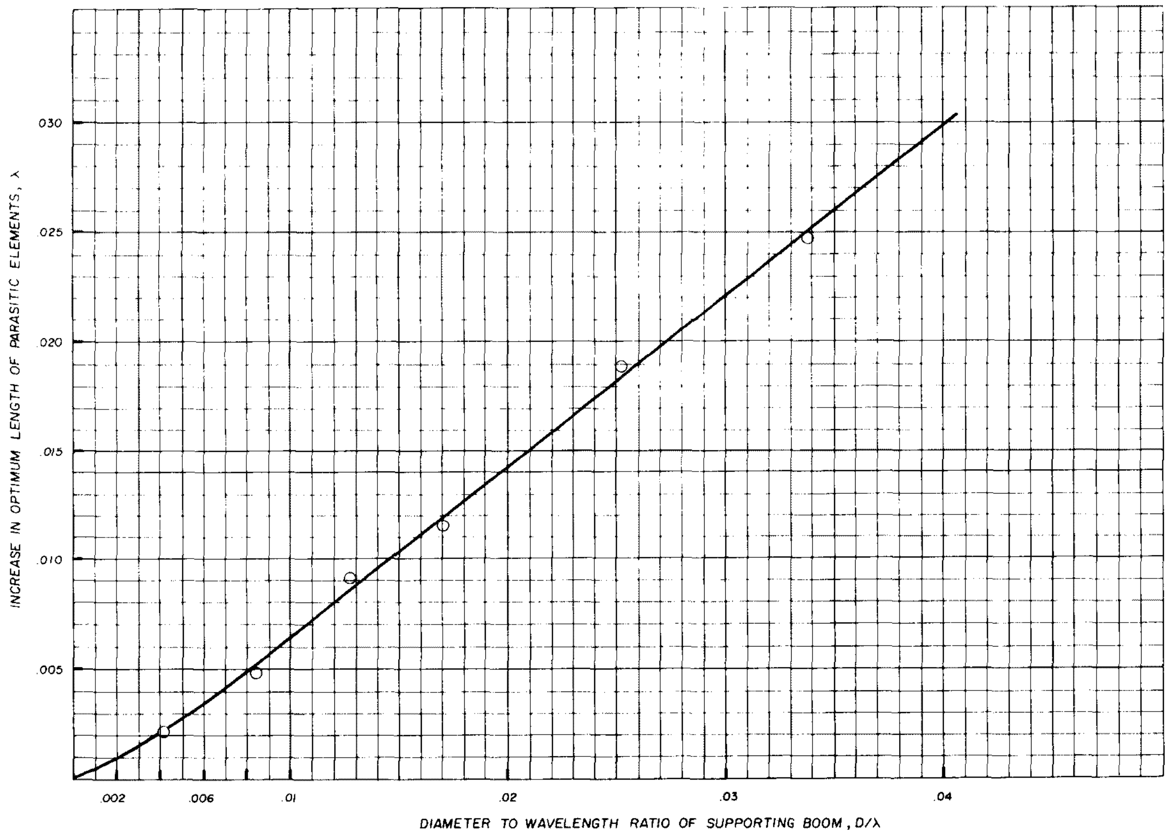


**fig. 5. Graph showing the effect of a supporting metal boom on the length of the parasitic elements.**

**Example 1.** It is desired to build a 6-meter Yagi with 10.2 dBd gain, using 0.5 inch (13mm) diameter elements mounted on insulating blocks above a 1.5 inch (38mm) diameter boom. This is the 1.2λ design in **table 1**.

The formula for wavelength is

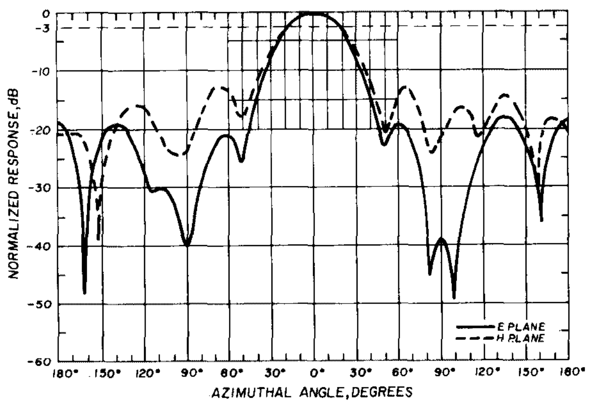


**fig. 6. Radiation patterns of a 6-element, 1.2λ long Yagi, built with the dimensions shown in table 1. Beamwidth of the $E$ plane is 40 degrees; $H$ plane beamwidth is 42 degrees.**

$$L = \frac{11803}{F} \text{ (inches)} \tag{1}$$

$$L = \frac{29980}{F} \text{ (cm)} \tag{2}$$

where $L$ = length
$F$ = frequency in MHz

| | |
|---|---|
| Frequency | 50.1 MHz |
| Wavelength | 235.6 inches (5.98 meters) |
| Element diameter (d/λ) | 0.0021λ |
| Reflector spacing | 47 inches or 120 cm (0.2λ) |
| Director spacings | 59 inches or 150 cm (0.25λ) |
| Boom diameter | not important, discussed later |
| Overall length | 283 inches (approximately 24 feet) or 7.2 meters (1.2λ) |

**1.** Plot the lengths of the parasitic elements for the 1.2λ design from **table 1** on the design nomograph

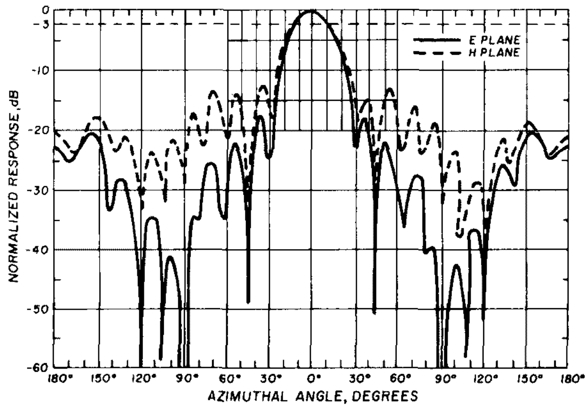


**fig. 7. Radiation patterns of a 15-element, 4.2λ long Yagi. Beamwidth of the** ***E*** **plane is 26 degrees;** ***H*** **plena beamwidth is 29 degrees.**

(see **fig. 8**) for parasitic elements with a diameter, $d/\lambda = 0.0085\lambda$.

$$L_R = 0.482\lambda$$
$$L_{D1} = L_{D4} = 0.428\lambda$$
$$L_{D2} = L_{D3} = 0.420\lambda$$

2. However, our element diameters are 0.0021λ so the element lengths must be adjusted. Draw a vertical line from 0.0021λ on the horizontal axis on the nomograph. This intersects the compensated lengths for the reflector and directors 1 and 4:

$$L_{R'} = 0.488\lambda$$
$$L_{D1'} = L_{D4'} = 0.451\lambda$$

**3.** Using a pair of dividers (or a compass), measure the distance between director 1 (D1) and director 2 (D2) determined in **step 1**. Transpose this distance from the point established in **step 2** to the left along the 1.2λ Yagi curve to 0.0021λ to determine the compensated length for directors 2 and 3:

$$L_{D2'} = L_{D3'} = 0.446\lambda$$

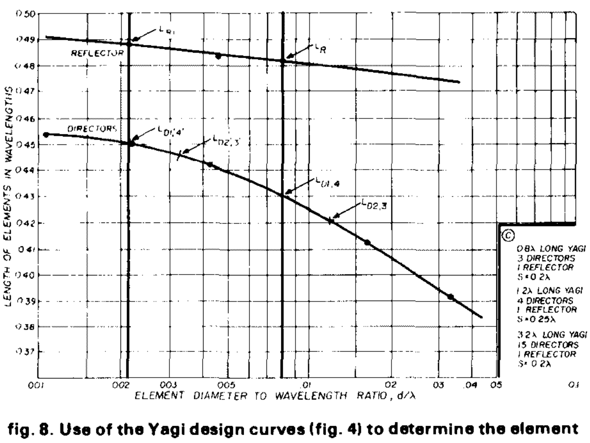


**fig. 8. Use of the Yagi design curves (fig. 4) to determine the element lengths for a 6-element, 50.1-MHz Yagi on a boom 1.2λ long (see example 1 in text).**

When I built this antenna I decided to use large element insulating blocks which I purchased from Swan Antennas (now KLM). Therefore, it wasn't necessary to put the elements through the boom. Since the wavelength is long with respect to the chosen boom diameter, I didn't feel that any boom correction was necessary. This was verified by subsequent tests. When the boom diameter represents a substantial portion of the operating wavelength, however, a correction for the boom diameter is required; this will be discussed in *example 3*.

The reflector and director lengths for the 50.1-MHz Yagi are as follows:

| | |
|---|---|
| Reflector | 0.488λ = 115 inches (2.92m) |
| Director 1 | 0.451λ = 106.25 inches (2.70m) |
| Director 2 | 0.446λ = 105.06 inches (2.67m) |
| Director 3 | 0.446λ = 105.06 inches (2.67m) |
| Director 4 | 0.451λ = 106.25 inches (2.70m) |

The approximate length of the driven element can be calculated from

$$L = \frac{5500}{F} \ (inches) \qquad (3)$$

$$L = \frac{13970}{F} \ (cm) \qquad (4)$$

where $L$ = length
$F$ = frequency in MHz

Therefore, at 50.1 MHz, the length of the driven element is 109.75 inches or 2.79 meters. For simplicity I decided to use a gamma match and to attach the driven element to the boom with a U bolt. During the matching adjustments the driven element was short-

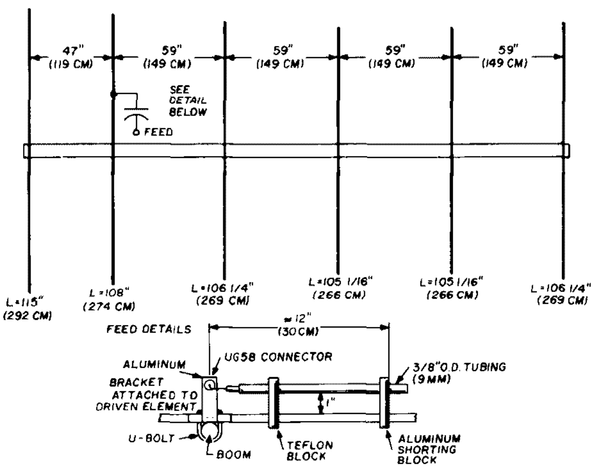


**fig. 9. Layout of a 6-element Yagi for 50.1 MHz on a 1.2λ boom. All elements are ½ inch (13mm) OD aluminum tubing, mounted on insulating blocks attached to a 1½ inch (38mm) OD aluminum boom. The gamma capacitor is approximately 12 inches (30cm) of RG-8/U coaxial cable with the outer jacket and shield removed, then inserted in a 3/8-inch (10mm) diameter tube.**

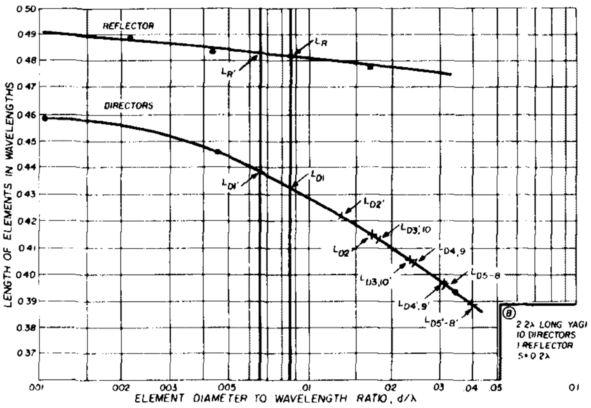


fig. 10. Use of the Yagi design curves (fig. 4) to determine the parasitic element lengths for a 12-element 205.25-MHz Yagi on a boom 2.2λ long (example 2).

ened to 108 inches (2.74m) for optimum vswr (the length of the driven element is not critical for maximum gain, as will be discussed later).

The completed 6-meter Yagi is shown in **fig. 9**. On-the-air receiving tests at W1JR have shown the 3 dB beamwidth to be between 40-45 degrees, while all sidelobes were at least 15 dB down; the front-to-back ratio was 18 dB. This agrees closely with the published NBS data.

**Example 2.** During the summer of 1973, when I was W6FZJ, transpacific tests to Hawaii were conducted on 220 and 432 MHz. Television video carriers seemed like a good propagation indicator so I designed a converter for Channel 12 on Mt. Haleakela on Maui. Since I had no good designs for a moderate gain Yagi with low sidelobes (to discriminate against Channel 12 TV stations in California), I chose the NBS 2.2λ Yagi design using 3/8 inch (1cm) diameter elements.

| | |
|---|---|
| Frequency | 205.25 MHz |
| Wavelength | 57.5 inches (1.46 meters) |
| Element diameter (d/λ) | 0.0065λ |
| Reflector spacings | 11.5 inches or 29.2 cm (0.2λ) |
| Director spacings | 11.5 inches or 29.2cm (0.2λ) |
| Boom diameter | not important, discussed later |
| Overall length | 126.5 inches or 3.21 meters (2.2λ) |

**1.** Plot the director element lengths for the 2.2λ Yagi design from **table 1** on the design nomograph (see **fig. 10**) for $d/\lambda = 0.0085$.

$L_R = 0.482\lambda$
$L_{D1} = 0.432\lambda$
$L_{D2} = 0.415\lambda$
$L_{D3} = L_{D10} = 0.407\lambda$
$L_{D4} = L_{D9} = 0.398\lambda$
$L_{D5}$ through $L_{D8} = 0.390\lambda$

**2.** Since the chosen element diameters are 0.0065λ, draw a vertical line from 0.0065λ on the horizontal on the nomograph. This intersects the compensated length for the reflector and the first detector:

$L_{R'} = 0.483\lambda$
$L_{D'} = 0.4375\lambda$

**3.** Using a pair of dividers, measure the distance be-

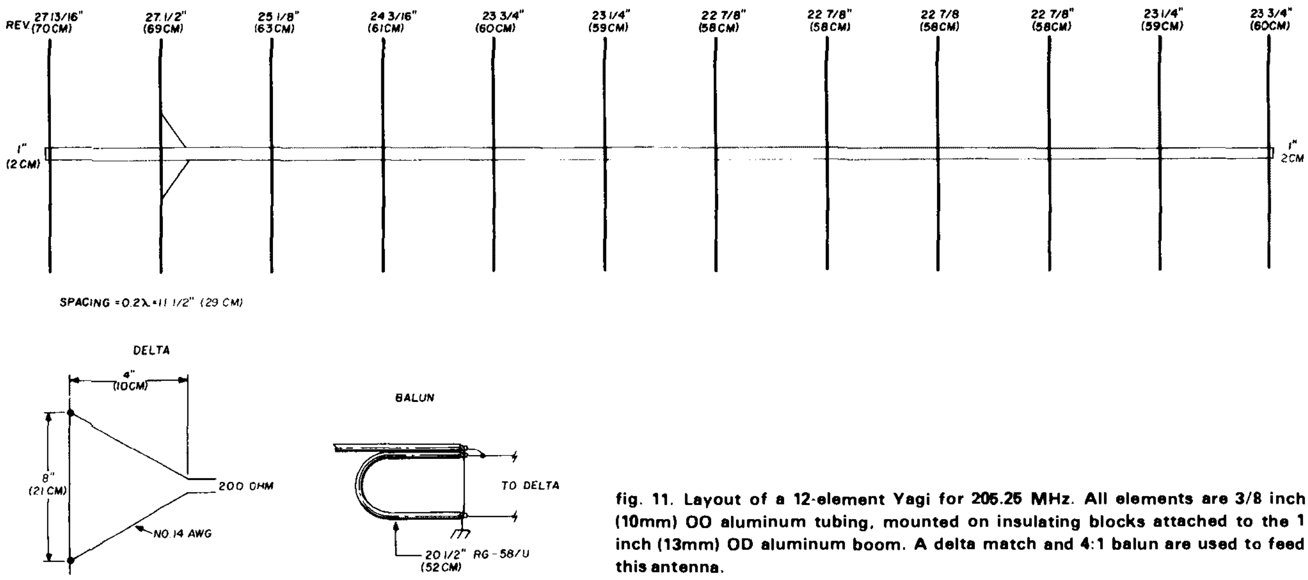


fig. 11. Layout of a 12-element Yagi for 205.25 MHz. All elements are 3/8 inch (10mm) OD aluminum tubing, mounted on insulating blocks attached to the 1 inch (13mm) OD aluminum boom. A delta match and 4:1 balun are used to feed this antenna.

tween director 1 (D1) and director 2 (D2) determined in step 2. Transpose this distance from the point established in step 2 to the left along the 2.2λ curve to determine the compensated length of director 2 ($L_D{}' = 0.421\lambda$). Now span the distance between directors 1 and 3 (D1 and D3) with the dividers, and move this dimension along the curve, making sure to reference D1′ (at the 0.0065 line). Follow this same procedure until all directors have been scaled. The remaining director lengths are as follows:

$$L_{D3}{}' = L_{D10}{}' = 0.414\lambda$$
$$L_{D4}{}' = L_{D9}{}' = 0.405\lambda$$
$$L_{D5}{}' \text{ through } L_{D8}{}' = 0.398\lambda$$

As in the case of the 6-meter Yagi, I decided to use element insulators which I purchased from KLM Electronics. Since the elements are mounted well above the boom, an element correction factor was not applied. The reflector and director lengths for the 205.25 MHz Yagi are as follows:

| | |
|---|---|
| Reflector | 0.483 = 27-13/16 inches (70.6cm) |
| Director 1 | 0.4375λ = 25-1/8 inches (63.9cm) |
| Director 2 | 0.421λ = 24-3/16 inches (61.5 cm) |
| Directors 3 and 10 | 0.414λ = 23-3/4 inches (60.5cm) |
| Directors 4 and 9 | 0.405λ = 23-1/4 inches (59.2cm) |
| Directors 5 - 8 | 0.398λ = 22-7/8 inches (58.1cm) |

You will note that these lengths have been slightly rounded off. The NBS report states that tolerances of 0.003λ should be maintained (0.173 inch or 4.4mm at 205.25 MHz). Furthermore, tests made by WØEYE

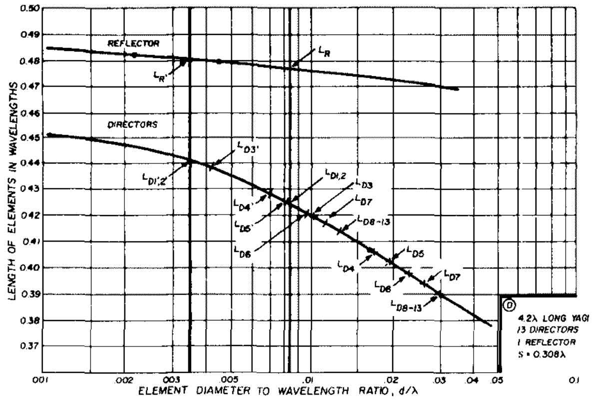


**fig. 12. Use of the Yagi design curves (fig. 4) to determine the parasitic element lengths for a 15-element Yagi for 432 MHz; boom is 4.2λ long (example 3).**

and W6FZJ in 1973 clearly showed that the gain and radiation pattern of a Yagi antenna degrades quite rapidly on the high side of the design frequency, but much more slowly on the low side. Therefore, if you must round off to a standard dimension, it is better to

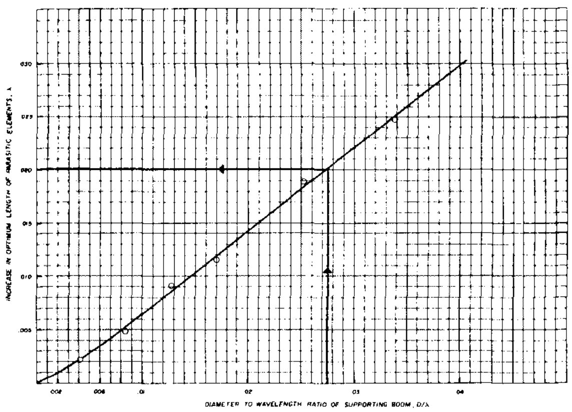

**fig. 13. Supporting boom correction factor for the 15-element 432-MHz Yagi. Boom diameter is 0.0275λ (3/4 inch or 1.9cm at 432 MHz); length of each parasitic element must be increased to 0.2λ.**

cut the director elements slightly shorter — not longer. Reflector length, on the other hand, should be rounded off on the long side. The element lengths for the Channel 12 Yagi were shortened to the nearest 1/16 inch, or only about 0.001λ.

For simplicity I decided to use a delta matching system and a 4:1 balun on this antenna. The driven element length was calculated using eq. 3. During final tests using the procedures outlined in reference 6 the driven element was extended slightly to obtain a 1:1 vswr. The length of the driven element is not a critical factor as long as the driven element is always shorter than the reflector.

The final design for the 205.25-MHz Yagi is shown in fig. 11. The desired discrimination to other Channel 12 television stations was achieved. This antenna is now in use at W1JR for indicating tropo, meteor shower, and aurora openings.

**Example 3.** The transpacific tropo tests mentioned in the previous example required an easily transportable antenna for the 432-MHz system to be installed at KH6BZF's station in Hawaii. I decided to use four 4.2λ Yagis similar to the WØEYE type[4] and stack them accordingly. The humidity and salt air are high in Hawaii so the elements were mounted through the boom using knurled 3/32-inch (2.4mm) diameter brass rods; this is similar to the method used on the W6FZJ extended, expanded collinear array described in *QST*.[7]

For the sake of brevity, steps 1, 2, and 3 will not be repeated here. However, the marked up nomograph for the 4.2λ 432-MHz Yagi is shown in fig. 13;

Since I decided to mount the elements through a metal boom, the elements must be lengthened to

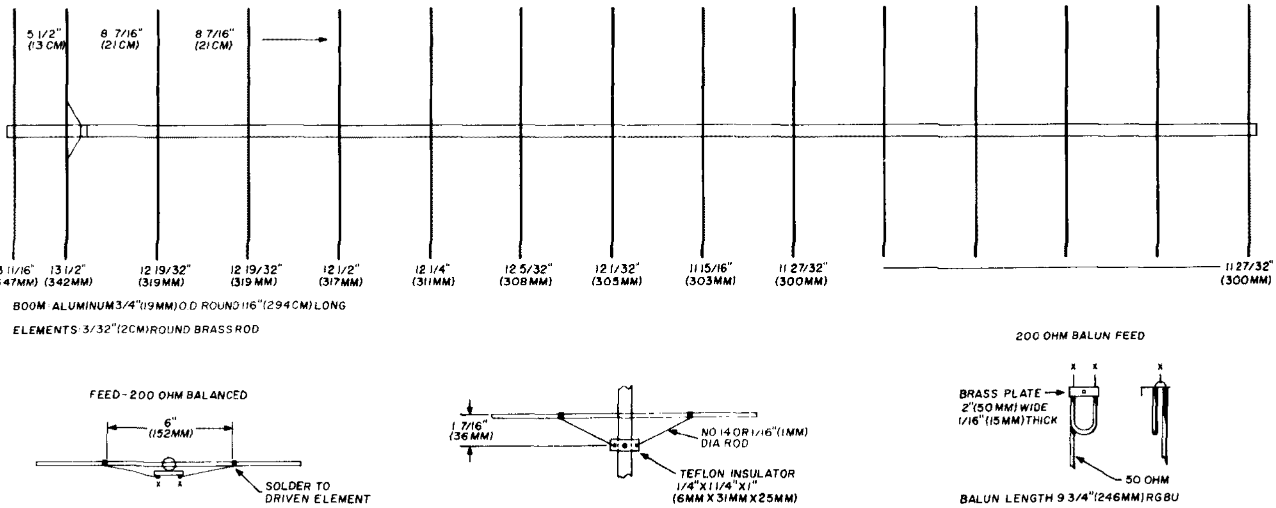


**fig. 14. Layout of the 15-element 432-MHz Yagi on a 4.2λ boom. All elements are 3/32 inch (2.4mm) OD brass rod; elements are knurled and tapped into under-size holes in the 3/4 inch (1.9mm) aluminum boom. Element spacing of all directors is 8-7/16 inches (21.4cm); reflector is 5½ inches (14cm) behind the driven element. Details of the delta matching system and 4:1 balun are also shown.**

compensate for the shortening effect of the boom.

| | |
|---|---|
| Frequency | 432 MHz |
| Wavelength | 27.32 inches (69.40 cm) |
| Element diameter (d/λ) | 0.00343λ |
| Reflector spacing | 5-1/2 inches or 13.9cm (0.2λ) |
| Director spacing | 8-7/16 inches or 21.4 cm (0.308λ) |
| Boom diameter | 3/4 inch or 1.9cm (0.0275λ) |
| Overall length | 115 inches or 2.915 meters (4.2λ) |

$L_{R'} = 0.480\lambda$
$L_{D1'} = L_{D2'} = 0.441\lambda$
$L_{D3'} = 0.438\lambda$
$L_{D4'} = 0.428\lambda$
$L_{D5'} = 0.425\lambda$
$L_{D6'} = 0.421\lambda$
$L_{D7'} = 0.417\lambda$
$L_{D8'}$ through $L_{D13'} = 0.414\lambda$

To determine the corrected element length, first convert the boom diameter (3/4 inch or 1.9cm in this case) to wavelength ($d/\lambda$) or approximately 0.0275λ. Draw a vertical line from 0.0275λ on the boom correction nomograph (see **fig. 13**) to the DATA line. Move to left-hand axis and read the correction factor; 0.02λ for this antenna. Add this length correction factor to all elements as shown below.

Note that all the director lengths have been rounded off to the *short* side, as in **example 2**. The driven element length was calculated with **eq. 3**, but a better match was obtained when it was extended to 13½ inches (34.3cm); a delta match with a 4:1 balun was used. The final design for the 432-MHz Yagi is shown in **fig. 14**.

This antenna stacks well at 1.6λ in the *H* plane, and 1.8λ in *E* plane. As tested at NBS, this quad Yagi array yielded 19.6 dBd. A one-way 432-MHz contact was attained between KH6BZF and W6FZJ in July, 1973 (don't ask me why it wasn't two-way because I'll cry loudly). During October, 1973, using only 200

| | | | |
|---|---|---|---|
| Reflector | 0.480 + 0.02 = 0.500 | 13-11/16 inches | (34.7cm) |
| Directors 1 and 2 | 0.441 + 0.02 = 0.461 | 12-19/32 inches | (32.0cm) |
| Director 3 | 0.438 + 0.02 = 0.458 | 12-1/2 inches | (31.8cm) |
| Director 4 | 0.428 + 0.02 = 0.448 | 12-1/4 inches | (31.1cm) |
| Director 5 | 0.425 + 0.v2 = 0.445 | 12-5/32 inches | (30.9cm) |
| Director 6 | 0.421 + 0.02 = 0.441 | 12-1/32 inches | (30.6cm) |
| Director 7 | 0.417 + 0.02 = 0.437 | 11-15/16 inches | (30.3cm) |
| Directors 8 - 13 | 0.414 + 0.02 = 0.434 | 11-27/32 inches | (30.1cm) |

watts and this array, EME signals from KH6BZF were copied and identified at W6FZJ. KH6BZF now uses this setup on Oscar 7, Mode B.

This article has presented a new and relatively precise way to consistently design and build Yagi antennas with optimum, reproducible gain characteristics — selecting a boomlength to suit your own requirements. Three design examples have shown Yagi antennas with demonstrated performance. If construction tolerances are held to 0.003λ maximum (0.001λ preferred), you should be able to design your own Yagis with the same excellent results. As pointed out earlier, director elements should be slightly shortened, while reflectors should be lengthened when rounding off the calculated dimensions.

Before actually starting to build a given design, double check your mathematics and scaling; it will pay off a 100-fold in time saved (and frustration). In those cases where the numbers in **table 1** do not agree exactly for the first director, reference at 0.0085 $(d/\lambda)$ on the chart. The feed methods are not critical, and attention to the details outlined in references 5 and 6 should fill any voids in this article.

In closing, I would especially like to thank Don Hilliard, WØPW (ex WØEYE), who first introduced me to this information, and to Peter Viezbickie who, after much prodding by Don and myself, finally published this wealth of information. Now you, too, can be an expert in designing your own Yagi antennas.

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**ham radio**

# the future of the amateur satellite service

During Phase III of the amateur satellite program AMSAT will place advanced communications satellites in high-altitude orbits which will allow long-range communications for up to 15 hours per day. This article discusses the capabilities of those satellites, and the financial support required from the amateur community

**Amateur radio stands** on the threshold of the most exciting and comprehensive change in its history, a change more revolutionary than that from spark to CW, or a-m to ssb, or the advent of vhf-fm repeaters. The Phase III Amateur Satellite Program, about which you'll be reading a great deal in the coming months, sounds more like science fiction than fact. However, in the past few years the *facts* have become increasingly clear: amateurs are already in command of the technology needed to produce a cost-effective satellite system — a *system*, not just a single satellite, capable of greatly enhancing the reliability of long-distance communications while simultaneously reducing the cost of the average amateur radio station.

One day, probably in late 1979, the first amateur Phase III satellite will be launched, and a new era in amateur communications will begin. It's possible that within ten years from today the majority of long-distance communications (over 50 miles or 80 km) by amateurs interested in DX, contests, traffic handling, and casual rag chewing will be by satellite. As a result, crowding of the high-frequency bands may be significantly reduced, even with a rapidly increasing amateur population.

Using satellite relays for global radio communications was first proposed by Arthur C. Clarke in the British journal *Wireless World* in 1945. Approximately 20 years later (March 9, 1965) the first active communications satellite, OSCAR 3, was launched. It may be hard to believe, but radio amateurs were communicating through OSCAR 3 months before the first commercial communications satellite, *Early Bird* (Intelsat I), was placed into orbit. Yet today, 12 years later, while satellites are carrying approximately two-thirds of all commercial transoceanic communications,[1] amateurs are still relying almost entirely on erratic high-frequency circuits for distant contacts.

Long-distance propagation on the high-frequency bands depends on signals being reflected by the ionosphere. A much more reliable communications system results when a satellite is substituted for the somewhat erratic ionosphere, and vhf or uhf bands are used for the radio links. You don't need to know much about the workings of the ionosphere to use the high-frequency bands; surprisingly, you don't need to know much about satellites to enjoy the advantages of this new mode of communications.

The satellite subsystem of primary interest to radio amateurs is the transponder, the electronic package which receives signals from stations on the ground and then retransmits them, on a different frequency with great amplification, back to earth. Although transponders are somewhat similar to 2-meter fm repeaters, there are significant differences: the linear transponders used on AMSAT satellites work equally well with ssb and CW signals, and they can simultaneously handle a large number of users.

To appreciate the communications capabilities which high-altitude spacecraft will provide we can

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compare communications links involving Phase III (high-altitude) satellites, Phase II (low-altitude) satellites such as OSCAR 6 and OSCAR 7 which are currently in orbit, and the 20-meter band. The comparison will consider a number of characteristics of specific interest to radio amateurs using these systems.

1. **Daily access time**. How many hours each day does the user have access to the satellite?
2. **Maximum communications range**. What is the maximum terrestial distance over which two stations can communicate?
3. **Communications performance**. How strong and intelligible are received signals? Can openings over specific paths be predicted reliably?
4. **Communications capacity**. How many stations can use the satellite at the same time?
5. **Frequencies**. What frequency bands will Phase III satellites use?
6. **Tracking techniques and operating schedules**. Will the paper work involved in tracking and checking operating schedules be complicated and laborious?
7. **Ground-station equipment**. How much transmitter power and how large an antenna will be needed? Will commercial or surplus equipment be available for a moderate cost ground station?
8. **Antenna aiming**. Will the direction in which the antenna is pointed have to be continually adjusted while operating?
9. **Miscellaneous**. How will factors such as satellite lifetime, signal time delay, lack of skip zone, Doppler shift, and crowding affect users?
10. **Financing the Phase III program**. Can amateurs afford the Phase III program?

## daily access time

The average amount of time that a satellite will be within range of a specific ground station each day (daily access time) is determined solely by the ground station's latitude. The first Phase III satellite will be injected into an orbit that initially places it within range of ground stations at mid-northern latitudes (this includes most of the United States) for about 15 hours each day, and within range of ground stations at mid-southern latitudes for about 5 hours each day. The first Phase III satellite will therefore provide northern hemisphere amateurs with as much access time as ten optimally spaced OSCAR 7 satellites! Can you recall the last time that 20 meters was open 15 hours per day on a regular basis?

As the years pass, ground stations will find that their average daily access time will change. By 1985 the first Phase III satellite will only be within range of northern hemisphere stations about 5 hours each day while southern hemisphere stations will have about 15 hours of access time each day. But don't despair, AMSAT is capable of producing two additional Phase III spacecraft before 1985. If these spacecraft are inserted into orbits similar to that of the first Phase III spacecraft, ground stations anywhere on earth will have access to at least one Phase III satellite for about 20 hours each day.

This article focuses on the potential impact of the amateur satellite program on amateur radio over the next ten years. The author, Dr. Martin Davidoff, K2UBC, is an assistant professor of mathematics at a community college in Maryland where he directs a National Science Foundation project involving satellites and college level science instruction. In conjunction with the NSF project, he recently authored a textbook featuring the AMSAT-OSCAR series of satellites, *Using Satellites in the Classroom: A Guide for Science Educators*. K2UBC obtained his doctorate in Physics from Syracuse University in 1974 and has held an amateur license since 1956. Editor.

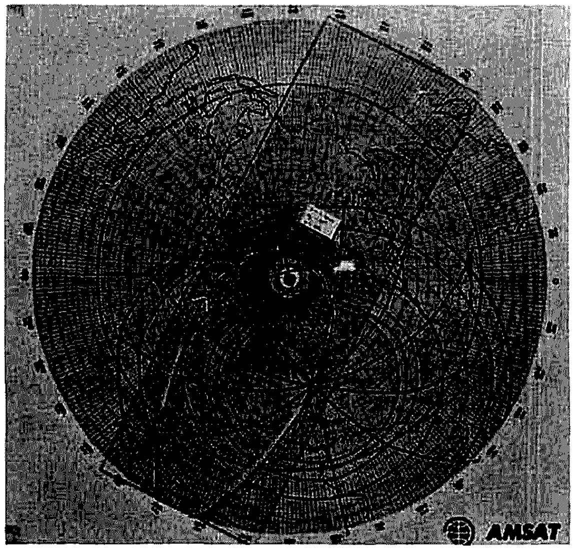



**fig. 1. Photograph of *Satellable* style tracking nomograph for elliptical orbit of the type used by the Phase III-A spacecraft.**

## maximum communications range

Phase II satellites provide a maximum communications range of about 5000 miles (8000km). While this is adequate for Worked All States and DXCC, it's not very satisfactory by high-frequency standards. Phase III satellites will enable amateurs to communicate over a much greater distance — up to about 11200 miles or 18000 km — leaving only a very

small region at the opposite side of the earth out of range. The 20-meter band will continue to reward its followers with somewhat unpredictable openings to all parts of the globe.

If you've had the opportunity of listening to ssb stations using the 432/146-MHz transponder on OSCAR 7, you know that satellites are capable of providing *telephone quality* links. The 20-meter band can provide similar performance, albeit in a somewhat erratic manner. As an example, assume it's August, 1977, and you're interested in the Denver-Frankfurt path during January, 1978. The best prediction high frequency propagation experts can offer is that 20 meters will probably be open over the path of interest sometime between 1600 and 1800 UTC on about 15 days during January. If the first Phase III satellite was in orbit at this time, you could predict with better than 99% certainty, that the Denver-Frankfurt circuit would be open for several hours every day in January during time slots specified almost to the minute.

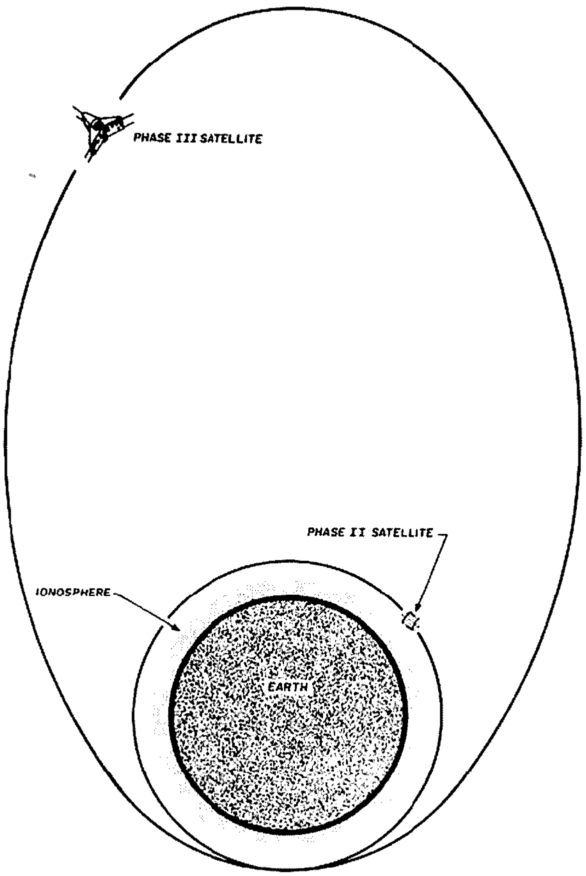



**fig. 2. Relative altitudes of Phase II satellites, Phase III satellites, and the ionosphere — shown approximately to scale. Drawing does not take inclination of orbits into account.**

The advantages of a Phase III satellite for prearranged point-to-point schedules are even more impressive when you consider a three-way contact between, for example, New York, London and Tokyo. Satellites will make such contacts possible daily with clocklike regularity. What are the odds of being able to accomplish this on 20-meters?

The received signal strengths observed by stations communicating through a high-altitude satellite will be largely independent of the distance between the stations. This results from the fact that the earth-satellite-earth distance and path loss are, for all practical purposes, constant regardless of how far apart the two ground stations might be. Since distance doesn't count, a station across town and one nearly halfway around the globe will produce similar signals if they are using similar equipment. In fact, listening to your own signal being returned from the transponder will indicate quite accurately how you sound to any station within range of the satellite.

## communications capacity

The first Phase III satellite will provide a band of frequencies, nominally 150 kHz wide, capable of handling hundreds of simultaneous conversations. All users will be sharing the 50-watt satellite transmitter. Therefore, ssb and CW will be the preferred modes because they efficiently use the available satellite power. The satellite transponders will also be able to handle slow-scan TV, RTTY, and fm. But these modes should only be used in emergencies, or for special experiments coordinated with AMSAT, because they use a disproportionate amount of satellite transmitter power and, in the case of fm, excessive bandwidth.

A recent cost-effectiveness study suggests that crowding will not become a serious problem with the first Phase III spacecraft until 30,000 users are equipped to transmit on the uplink frequency.[2] To prevent crowding problems AMSAT plans an ongoing Phase III construction program which is designed to keep pace with a rapidly increasing user population. Placing an additional satellite in orbit every 24 months appears to be a realistic goal.

## frequencies

The first Phase III satellite will include two transponders. Therefore, even if one transponder fails, the satellite will still be available for communication on a full-time basis. If possible, the two transponders on the first Phase III spacecraft will use reciprocal frequency combinations: one transponder will receive on 435 MHz and transmit on 146 MHz, the other will receive on 146 MHz and transmit on 435 MHz. Users will be able to compare the performance of both transponders and express their preferences for scheduling and for future satellites.

The amateur satellite program will continue to rely heavily on the 146-MHz and 435-MHz bands

throughout the 1980s. In the mid or late 1980s Phase III satellites are likely to include links at even higher frequencies such as the 920 MHz (32cm) and 2.3 GHz (13cm) bands; specific plans must await the outcome of the 1979 World Administrative Radio Conference.

Low-altitude (Phase II) satellites may continue to use 10 meter downlinks, a band which is not suitable for Phase III. Readers interested in the factors involved in selecting frequencies for amateur satellite systems are referred to the excellent paper by Ray Soifer, W2RS.[3]

## tracking techniques and operating schedules

You may be pleasantly surprised to learn that you won't need to know anything about tracking to use a Phase III satellite. After the first Phase III spacecraft is in orbit, you'll be able to turn on your receiver (with an omni-directional antenna connected) and check to see whether or not the band is open (satellite within range) by simply tuning for signals. About 65% of the time stations in mid-northern latitudes (most of the United States) will find that they're in luck — signals will be present. If a second Phase III spacecraft is launched into a similar orbit, the probability of finding the band open will be about 90%. When the band is open, you'll switch to a beam antenna and home in on the satellite by peaking your S-meter on a beacon signal. The same antenna setting will work for all stations received via the satellite. Peaking the antenna every 15 minutes should be more than sufficient.

While the casual user can get away without any knowledge of tracking, some tracking skill (like a little insight into 20-meter propagation) will pay big dividends by enabling you to predict specific openings, to rare countries.

Many radio amateurs think that satellite tracking is very difficult and requires a strong mathematical aptitude. This just isn't true. Tracking is a simple, mechanical skill that takes only a few minutes to learn, and the only math needed is basic arithmetic. The ability to predict 20-meter propagation stands in sharp contrast; it's an impressive skill which requires a great deal of knowledge and experience.

Most tracking methods use some sort of nomograph which usually consists of a map and transparent overlay. Until recently everyone had to build their own tracking nomograph from scratch — a straightforward but tedious job which is no longer necessary since excellent commercial tracking aids are now available.[4] With minor modifications the basic tracking techniques and nomographs used with OSCAR 6 and 7 will also work for the radically different orbits which early Phase III spacecraft will introduce. In fact, these same nomographs were actually used to evaluate the communications capabilities provided by the various orbits considered for Phase III. Construction details for Phase III tracking nomographs will be published in the near future.

In the past, OSCAR 6 and 7 users have sometimes complained about the bookkeeping involved in determining which OSCAR 6 orbits are available for general use and which OSCAR 7 transponder is scheduled to operate. The latest W6PAJ orbit calendar[5] eliminates most of the bookkeeping drudgery by clearly listing the times and operating status for every OSCAR 6 and OSCAR 7 orbit during 1977. AMSAT will have a great deal of flexibility in scheduling future Phase III satellites because they will be controlled by onboard microcomputers that can be programmed by suitably equipped ground stations. User convenience will be the primary consideration when satellite schedules are chosen so bookkeeping requirements should be minimal. Tracking nomographs and orbit calendars will be made available for Phase III satellites soon after they're in orbit.

## ground station equipment

**Receiving.** Ground stations working with Phase III satellites will need a good ssb/CW receiver capable of tuning a few hundred kHz around 146 MHz and/or 435 MHz.

**Transmitting.** A CW or ssb transmitter with about 50 watts output at 146 or 435 MHz will be required for the uplink.

**Antennas.** Ground station communication via the Phase III satellite will usually require moderate gain (10-15 dBi) beam antennas for receiving and transmitting. A typical antenna array may consist of two or more Yagis mounted on a common mast using a single set of azimuth and elevation rotators. The entire structure can be smaller and lighter than the average three-element beam used on 20 meters.

The selected antenna site should place the antenna clear of surrounding objects and relatively close to the operating position since feedline losses are an important consideration at 146 and 435 MHz. As a result, a chimney mount will often be as effective as a large tower. Neighbors (and zoning committees) will probably be unable to distinguish between a roof-mounted Yagi array for satellite work and a large television antenna!

Although beam antennas will usually be required for reliable communications, simple omnidirectional antennas will also be useful at times. For example, an omnidirectional receiving antenna can be used during the entire orbit to determine whether the satellite is within range. In addition, omnidirectional transmitting and receiving antennas will sometimes be convenient for communication when the satellite altitude is relatively low (less than 15% of each orbit).

**General**. The satellite ground station that you put together will no doubt depend on the size of your pocketbook, the equipment you already own, the amount of time you have to devote to the project, and the transponder frequency.

Here are some options that you may want to consider. If you presently own a good high-frequency receiver, a top-line vhf or uhf converter will provide you with a state-of-the-art receiving setup. For transmitting, the 10-watt multimode or ssb/CW transceivers currently available for 2 meters and 70 cm look like a good choice. A linear amplifier with 6-10 dB of gain will keep the transmitting antenna requirements within modest limits.

Numerous pieces of commercial equipment suitable for satellite work (converters, transmitters, antennas) are currently available off the shelf; I'm not speculating as to what the future may offer. If you have some time and a little technical knowhow, you'll be able to put together a relatively inexpensive station using a surplus fm strip as a CW transmitter and homebrew helix antennas. In any event, if you're currently thinking of investing in a vhf or uhf fm transceiver or amplifier, consider paying a little extra to obtain a rig with ssb/CW capabilities and purchasing an amplifier that can be run in the linear mode for ssb.

Let's look at the equipment procurement problem from a different perspective by putting ourselves in the shoes of a newcomer to amateur radio five years from now (1982). If the newcomer intends to stick with the hobby for quite a few years and wants to set up a first-class station for local and DX work with new, off-the-shelf equipment what are the options?

**Option A**

Synthesized 2-meter fm transceiver

Separate transmitter and receiver for high-frequency bands

Kilowatt high-frequency amplifier

50-foot (80m) tower and triband Yagi

**Option B**

10-watt multimode 2-meter transceiver

10-watt multimode 70-cm transceiver

50-watt, 2-meter and 70-cm linear amplifiers

Modest roof-mounted antenna array

While each of these options will provide roughly equivalent capabilities, **Option B**, which depends on satellites for long-distance work, costs approximately half as much as **Option A**. Since prices for vhf and uhf ssb/CW gear are likely to decrease when a big new market opens up, the financial advantage of **Option B** is likely to increase.

The first Phase III satellite will be a moving target. The question that concerns radio amateurs is: How difficult will it be to track this satellite with a moderate gain beam? In other words, how frequently will the ground station operator be required to adjust the azimuth and elevation controls? The answer depends on a number of factors including: satellite orbit characteristics, location of the ground station with respect to the satellite, and the beamwidth of the antenna.

An analysis of the problem, taking these factors into account, shows that a ground station using a moderate gain beam will, on the average, need to adjust azimuth and elevation controls about once every 15 minutes during most of the orbit. However, there will be times while the satellite is near the low point on its orbit (perigee) when almost continual adjustment of beam elevation and azimuth will be required. Since signals will be very strong near perigee, ground stations will find it convenient to switch from beams to simple fixed omnidirectional antennas during this relatively short period of time.

Let's compare the dynamic antenna aiming requirements for Phase III, as just outlined, to requirements for Phase II satellites and the 20-meter band. Radio amateurs who have been using low or moderate gain beams to access OSCAR 6 and OSCAR 7 will find that they'll be able to pay far less attention to azimuth and elevation controls when they communicate via a Phase III satellite. Operators familiar with 20 meters will probably also be pleased to observe how a single antenna setting will work for all stations using the satellite; there's no need to repeatedly adjust the antenna for each weak DX signal.

AMSAT hopes that some future Phase III satellites may be placed in geostationary (or nearly geostationary) orbits.[6] A satellite in a geostationary orbit will appear to remain fixed directly above a spot on the earth's equator; a satellite in a nearly geostationary orbit will appear to drift slowly in longitude while remaining directly above the equator. Ground stations using these satellites will only need to adjust azimuth and elevation controls when switching from one satellite to another or when turning on the ground station after it's been off for a day or longer.

## miscellaneous

**Lifetime**. Satellite lifetime concerns radio amateurs for several reasons. First, lifetime affects the yearly cost of the satellite. This subject will be covered in detail in the next section. Second, lifetime affects the long-term reliability of a satellite communications system. If a system depends on a single satellite, satellite failure shuts down the system. Potential

users of a system based on a single satellite are naturally hesitant about investing time, energy, and money in a ground station that might suddenly have no function. Although the long lifetimes of OSCAR 6 and OSCAR 7 have alleviated this concern to some extent, the real solution is to produce a multiple satellite system so that the failure of a single satellite causes users only minor inconvenience. The term Phase III connotes just such a system. For this reason amateurs building ground stations for Phase III need not worry about their station suddenly becoming useless.

Let's look briefly at some of the plans for implementing the Phase III system. Experience with Phase II has shown that it's reasonable to expect operational lifetimes of five years for Phase III satellites. During the five year period following the launch of the first Phase III satellite, additional spacecraft will be placed into orbit; a new satellite every two years is a realistic goal for the 1977-1985 time frame. By 1984 the system should average three or more Phase III spacecraft in orbit and operating at any given time.

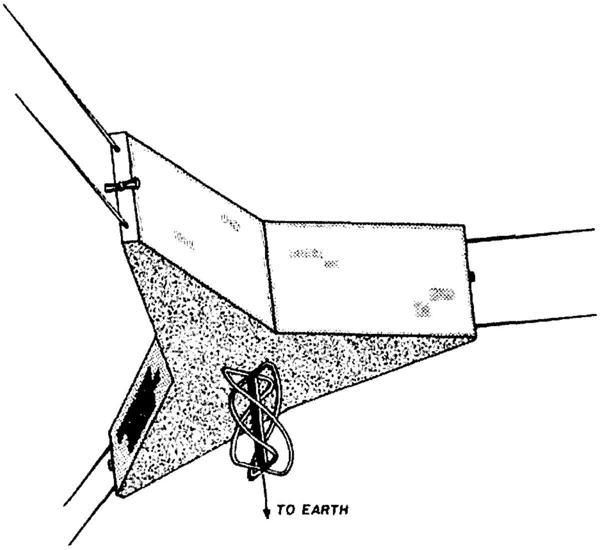
In the past amateurs thought that a satellite's useful life ended when it ceased to function. In the future AMSAT might decide to retire an old operating spacecraft from service in order to replace it with a new, more powerful model before total failure occurs

**Time delay**. Time delays from 10 to 300 milliseconds on the earth-satellite-earth path will make continuous monitoring of your own downlink distracting — to say the least. This is just one illustration of the many subtle differences between terrestial and satellite communications systems which amateurs will encounter in the future.

Each time amateurs have introduced new communications systems (ssb in the 1950s, for example, or fm repeaters in the 1960s), they've had to develop new operating procedures. Satellite systems will also require such innovations. One way to compensate for the time delays encountered while using high-altitude satellites might be to set the hang time on the vox or CW break-in system to 300 milliseconds and pause periodically for a second or two to enable the other party to break in.

**No skip zone**. Satellite communication systems do not exhibit skip zones. Consequently, it's easy to tell if a frequency is being used and a lot of unintentional interference can be avoided. In addition, round-table and net operation will be greatly facilitated since all users will hear each other. No problems to cure here — just a big bonus for satellite operators.

**Doppler shift**. Anyone who has listened to signals from a low-altitude satellite such as OSCAR 6 or 7 has probably noticed the pronounced downward shift in frequency that signals exhibit during nearby passes. A similar downward shift in frequency can be observed when a train passes with its whistle blowing. In both cases the frequency shift (called Doppler shift) is observed even though the source frequency is constant. The magnitude of the Doppler shift which amateurs encounter during satellite com-




**Sketch of AMSAT-Phase III-A spacecraft.**

munications depends on the relative velocity between satellite and ground station and the frequency being used — higher relative velocities and higher frequencies produce greater shifts.

Single-sideband communication is especially sensitive to Doppler shift since frequency changes of a few hundred cycles can make an ssb signal unintelligible. The largest Doppler shift that amateur radio operators have so far encountered during two-way communications occurs when OSCAR 7 passes directly overhead with the 432/146 MHz transponder in use. Under these conditions the Doppler shift is annoying, but ssb stations are able to compensate by frequent receiver tuning.

The first Phase III satellite will be moving slowly (relative to the surface of the earth) most of the time. During this portion of the orbit, Doppler shift will be smaller than observed with Phase II spacecraft which use the same transponder frequency combinations. However, there will be a small segment of each orbit, amounting to less than 10% of the period, when Doppler shift may be annoying, although ssb communications should still be possible.

Later Phase III satellites may be placed in geostationary orbits. Since these satellites will appear to re-

main fixed in space (no relative motion between satellite and ground station), no Doppler shift will be observed. In sum, Doppler shift will be of minor concern only with early Phase III satellites and of no concern with geostationary Phase III satellites.

**Crowding**. It has been estimated that the first Phase III satellite can accommodate 30,000 users equipped to transmit on the uplink bands. The estimate is based on ssb stations using 100 kHz and CW stations using 50 kHz of the transponder. If crowding becomes a problem before the follow-on Phase III spacecraft is launched, users have several options. Many may shift from ssb to CW to accommodate more stations in a given bandwidth. However, the opposite strategy, switching from CW to ssb, may actually be more effective in reducing crowding problems because a station can pass a given amount of information in a much shorter time period with ssb than with CW, while using less total spacecraft energy. This strategy would work only if amateurs limit themselves to essential information — a questionable objective.

Crowding effects can also be minimized by increasing the amount of roundtable and net operation. Phase III satellites are especially suitable for such use since there will not be any skip zone, and Doppler shift will be minimal. In any event, it should be evident that a number of viable options exist in response to any crowding problems that may temporarily occur. I have no doubt that amateurs who were raised on the 40- and 80-meter Novice bands will be able to devise satisfactory solutions.

## financing

Phase III will become a reality only if the international amateur community is willing and able to financially support the program. While large donations from individuals, corporations, and foundations are needed to produce the first Phase III spacecraft, a long-term Phase III program depends on small donations from a very wide base of support in the amateur community.

An educated guess places the procurement cost for a commercially built Phase III satellite in excess of five million dollars. An early AMSAT estimate pegs the cost of the first Phase III satellite at two-hundred thousand dollars, a considerably smaller but still imposing figure. A much more meaningful number is the cost per year of service. Experience has shown that it's reasonable to expect an operational lifetime of five years for a Phase III spacecraft, so the cost per year of service for the first Phase III satellite is expected to be less than $45,000.

Let's look at this figure more closely. When the number of amateur radio operators equipped for the uplink reaches 15,000 (half the estimated spacecraft capacity), the yearly cost per user will be less than three dollars! This means that when AMSAT membership reaches 50% of user capacity, the current $10 AMSAT membership fee* should be able to support an expanding program of satellite construction and provide for membership services, educational programs, and the *AMSAT Newsletter*. However, AMSAT satellites will always be free access and available to anyone licensed to operate on the uplink frequencies. Readers interested in cost breakdowns for the first Phase III satellite should read the current series in *QST* by Jan King, spacecraft project manager.

## general comments

Phase III satellites have been compared to Phase II satellites and 20 meters throughout this article. The points of comparison were chosen to illustrate Phase III satellite capabilities in familiar terms. As a result many of the unique and desirable characteristics of Phase II satellites and 20 meters have been ignored. I will now briefly discuss some of these features.

The 20-meter band will certainly remain a favorite of amateurs for a number of reasons. It's probabilistic nature is actually a very appealing characteristic — a skilled, knowledgeable, patient operator with a simple low-power 20-meter station will eventually be rewarded with exciting openings to the entire world. In addition, RTTY and sstv buffs can use 20 meters for hour after hour without ever worrying about using an unfair amount of satellite power.

Low-altitude satellites can be used by very simple ground stations. Contacts through Phase II satellites have been made with as little as 100 milliwatts, and numerous amateurs have had contacts using less than 1 watt of transmitter power; it is therefore possible to communicate through low-altitude satellites using small hand-held portable units. Low-altitude satellites can also be used in a broadcast mode, for example, to carry a single bulletin to the entire United States via 2-meter fm. Because of these features AMSAT will launch another Phase II satellite in late 1977 and continue the Phase II program through the 1980s. If you haven't already done so, try your hand at using the low-altitude satellites currently in orbit; they can provide a great deal of fun and excitement.

I think it's clear that Phase III will add a new dimension to amateur radio by augmenting the existing long-distance communication modes, not by replacing them.

*Regular AMSAT membership is $10 per year; life membership is $100. *Write to AMSAT, Post Office Box 27, Washington, DC, 20044.*

With the uncertain outcome of amateur frequency allocations at the World Administrative Radio Conference in 1979, and the rapidly increasing amateur population in the United States and abroad, the question is no longer "Can we afford to go ahead with the Phase III project?" The question is, "Can we afford not to go ahead with the Phase III project?"

The European Space Agency has selected the first AMSAT Phase III satellite for a late 1979 launch. The selection was a significant honor for the AMSAT team, but satellite construction can proceed on schedule only if AMSAT can obtain adequate funds. The money needed can be raised if amateurs are willing to demonstrate their commitment to the Phase III satellite program by joining AMSAT now, before the first Phase III satellite is launched.

Individuals who would like to make a more substantial contribution are encouraged to do so by donating money — contributions are tax deductible under section 170 of the IRS codes and/or donating time — volunteers are needed for a myriad of Phase III related activities (and you don't have to live in the greater Washington, D.C. area to participate).

Ten years from today amateurs will probably look back at the years 1977-1981, bracketing the launch of the first Phase III satellite, as one of the most exciting periods in the histoory of amateur radio. Take part in making history and enjoy it as it happens; *invest yourself in the future of amateur radio.*

## acknowledgements

Much of the information presented in this article was derived from conversations with Jan King and Perry Klein of AMSAT. Their comments on prepublication drafts of this article were extremely useful. I am also grateful to Linda Davidoff. Her editorial assistance considerably improved the clarity of the manuscript.

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**ham radio**

# homebrew
# Touch-Tone encoder

Details for building a simple encoder designed around the 555 timer IC

**For those who like to make their own** *Touch-Tone*† encoders, a cursory review of the amateur radio literature shows several reliable circuits. Several are built around the Western Electric Model 35 pad.[1-4] Another encoder used a pair of 565 IC voltage-controlled oscillators,[4] while others have used the recently developed MC14410 cmos encoder.[5,6] However, to my knowledge, no encoder has been built or described using the type 555 IC timer.

This article is the result of the challenge to design and build a simple *Touch-Tone* encoder with automatic PTT control using the 555 timer.

## design

For a 12-button pad, *Touch-Tone* information is encoded in pairs using two of seven possible frequency tones. As shown in **table 1**, these seven tones are divided into a "low" group (rows) of 697, 770, 852, and 941 Hz; and the "high" group (columns) of 1209, 1336, and 1477 Hz.

To generate the required *Touch-Tone* codes, two 555 timers are required, each connected as an astable multivibrator with an output frequency given by

$$f(Hz) = \frac{1}{0.693(R_A + 2R_B)C}$$

$R_A$ is the resistance between the timer discharge output and $+V_{CC}$, and $R_B$ is the resistance between the threshold input and discharge output. As shown in **fig. 1**, $R_A$ is replaced by a resistive divider string for both the low- and high-tone oscillators.

**table 1. Frequencies used in the Touch-Tone signaling system.**

| | high-tone group | | |
|---|---|---|---|
| **low-tone group** | **1209 Hz** | **1336 Hz** | **1477 Hz** |
| 697 Hz | 1 | 2 | 3 |
| 770 Hz | 4 | 5 | 6 |
| 852 Hz | 7 | 8 | 9 |
| 941 Hz | * | 0 | # |

For the low tone oscillator, U1, letting $R_A$ = R1 = 4.3k ohms, C = 0.047 $\mu$F, and $f$ = 941 Hz, solving for $R_B$ yields 14,164 ohms. To generate the next lower tone (852 Hz), $R_A$ is now equal to R1 + R2, so that R2 = 3.3k ohms. For the 770-Hz tone, $R_A$ now equals R1 + R2 + R3, giving R3 = 3.9k ohms. In a like manner, R4 = 4.3k ohms.

The high-tone oscillator U2, is designed in a similar manner. Starting with the 1477-Hz tone, letting R5 = $R_A$ = 3.9k ohms and $C$ = 0.047 $\mu$F, R6 = 2.2k ohms and R7 = 2.4k ohms.

For both oscillators the outputs are taken from the

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†*Touch-Tone* is the registered trademark of the American Telephone and Telegraph Company.

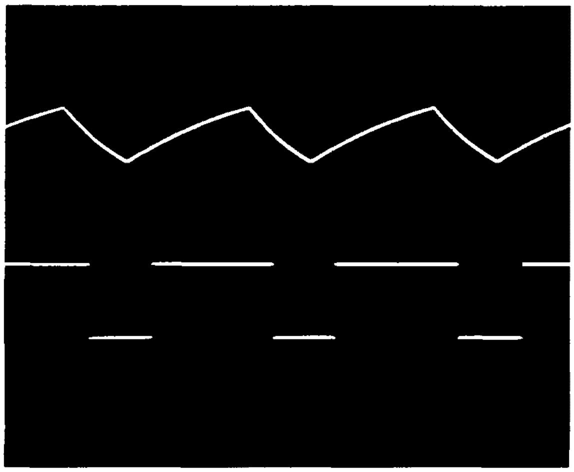

fig. 2. Type 555 timer output waveforms. When connected as an astable multivibrator upper trace is obtained from pins 2 and 6; lower trace from pin 3.

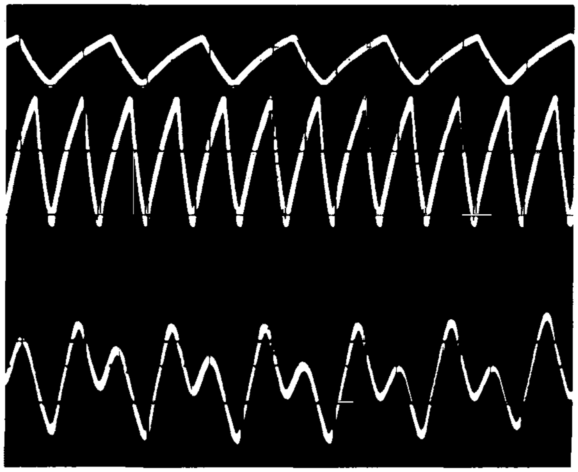

fig. 3. Addition to the low-tone oscillator (top trace) and the high-tone oscillator (center trace) to give a two-tone output signal when the digit 2 is pressed.

timer discharge junction and trigger pins (pins 2 and 6), which produce a pseudo-triangular waveform between 1/3 and 2/3 $V_{CC}$ (**fig. 2**). A 741C op-amp, U3, adds the output of both oscillators, shown in **fig. 3**, and is coupled to a 10-k ohm pot, which is an output-level control.

The automatic PTT control consists of another 555 timer, U4, connected as a 1-second one-shot and relay K1.

## components

For good thermal stability the 0.047-$\mu$F capacitors should be either tantalum or mylar, and resistors R1-R9 should be 1%. Several manufacturers currently advertise a 4X3 pad similar to the Chomerics type ER-21623. R10 and R11 are 10-turn pots. If desired, a type 556 dual timer can replace U1 and U2. **Fig. 4** compares the pin connections to of the 555 and 556 timers.

## adjustment

Start by pressing the * key and adjust R10 so that the low-group oscillator reads 941 Hz at pin 3 of U1. Consequently, frequencies of 852, 770, and 697 Hz should be obtained to within 2% when the numbers 7, 4, and 1 are respectively pressed. For the high tone

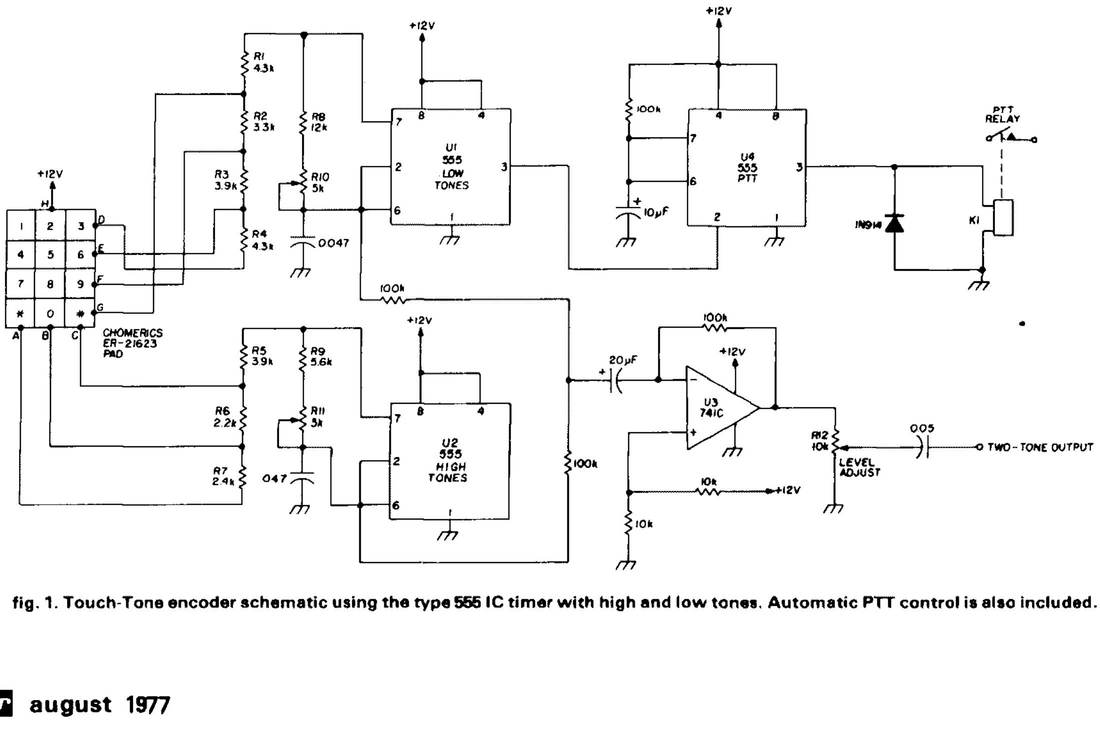


fig. 1. Touch-Tone encoder schematic using the type 555 IC timer with high and low tones. Automatic PTT control is also included.

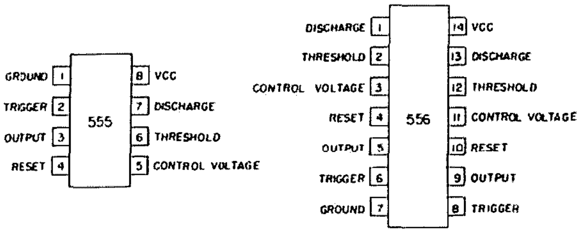


**fig. 4. Pin-connection comparison for the type 555 and 556 timers.**

group, press the # key and adjust R11 so that the oscillator reads 1477 Hz at pin 3 of U2. Consequently, frequencies of 1336 and 1209 Hz should be obtained to within 2% when the **0** and * keys are pressed.

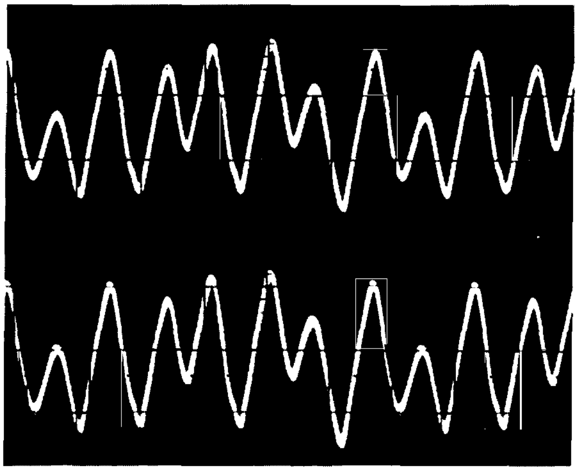

**fig. 5. Comparison of two-tone output obtained from the 555 timer encoder (upper trace) with Western Electric Model 35 pad (lower trace) for the digit 1.**

A comparison of the digit **1** generated by a Model 35 pad and the 555 encoder is shown in **fig. 5**.

For proper operation output-level control R12 should be set against a deviation meter; otherwise on-the-air testing will be required to set the output level.

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**ham radio**

# frequency-marker standard

## using CMOS logic

Low power consumption and fast switching speeds through the hf bands are featured in this circuit

A few frequency-marker standards have appeared in the amateur literature, but most use linear ICs or TTL logic. The frequency standard described here uses CMOS CD4000-series logic elements, which result in reduced power consumption but switch fast enough to allow sufficient switching speed and harmonic energy throughout the hf bands for good response.

### circuit description

The divider arrangement is shown in **fig. 1**, in which a 400-kHz crystal-reference frequency is used. A 1-MHz reference would be better; however, the 400-kHz frequency was chosen because of available parts.

The oscillator is somewhat different than others using logic elements,[1] such as that using TTL ICs.[2] Resistor **R1** keeps the input of the first NOR gate, U1A, stable; otherwise the oscillator refuses to oscillate at the crystal frequency. Resistor R2 is not critical and can be any value between 1 and 20 megohms. It should be close to 10 megohms if temperature extremes are experienced.[1] Resistor R3 provides load isolation between Y1 and the NOR-gate output. RC network C2, R3, R2, C1 forms a feedback arrangement similar to that recommended by reference 1, wherein C2 can be adjusted to vary crystal frequency to zero-beat with WWV or CHU. Capacitor C1 is used for crystal loading and centering of the crystal-frequency range.

Buffer NOR gate U1B isolates the crystal oscillator from the following divider chain, U2-U6. U2A-U3B are a divide-by-4 circuit in which the logic elements are JK flip-flops connected as D flip-flops (D flip-flops could be used here).

Calibration markers were desired at 200, 100, 50, 25, 10 and 5 kHz. This means that some additional dividing was necessary along with the following flip-flops to allow the desired symmetrical waveshape, since any feed back-type counter-divider produces an unsymmetrical waveform output. Therefore, U4 was chosen as a divide-by-N counter IC, and the latch arrangement of U5 was used to reset the selected divide-by-5 logic.

The 100-kHz output was first divided by 5, which was then divided by 2 to provide the symmetrical 10-kHz output; this output was then divided by 2 to provide 5-kHz. A switch, S1, and a final buffer, U1C, were used between selected output and the output coupling point. (A small 100-pF capacitor can be used for output coupling.)

### construction

The photo shows component layout. I used perf board with straight pin-to-pin wiring. If you wish to use a PC board, one could be easily designed. Nothing is critical about parts layout; however, short leads should be used in the oscillator circuit to preclude problems with stray capacitance.

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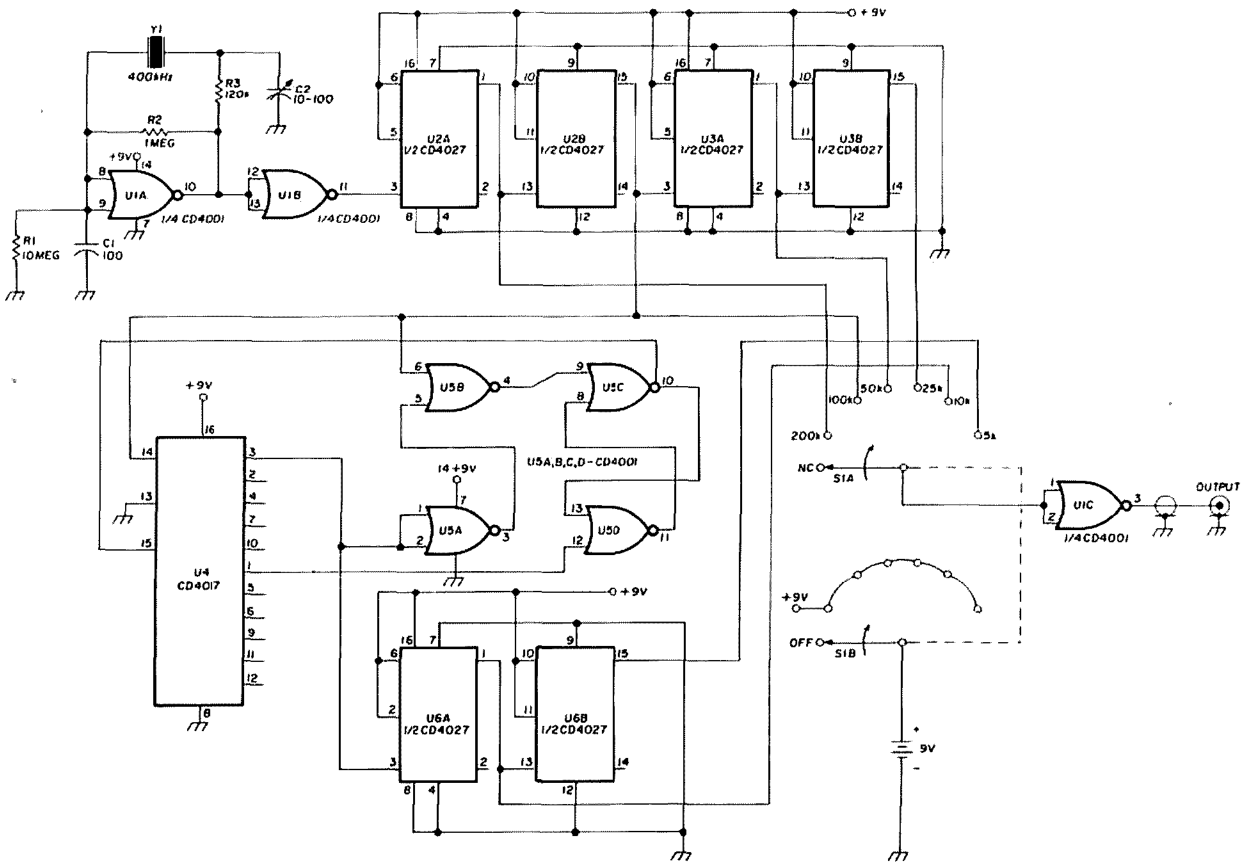
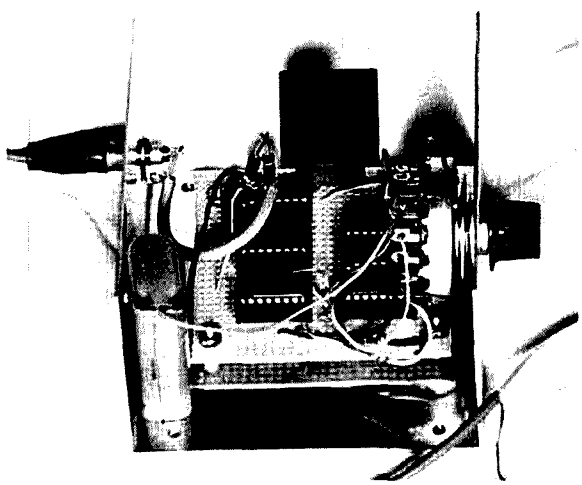



fig. 1. The CMOS IC frequency-standard schematic. Type JK flip-flops connected in a D configuration are used in the divide-by-4 circuit (U2A-U3B). Type D flip-flops could be used here.



Component arrangement of the CMOS frequency standard. Layout is not critical; straight pin-to-pin wiring is used. Oscillator requires short leads for minimum frequency variation due to stray capacitance.

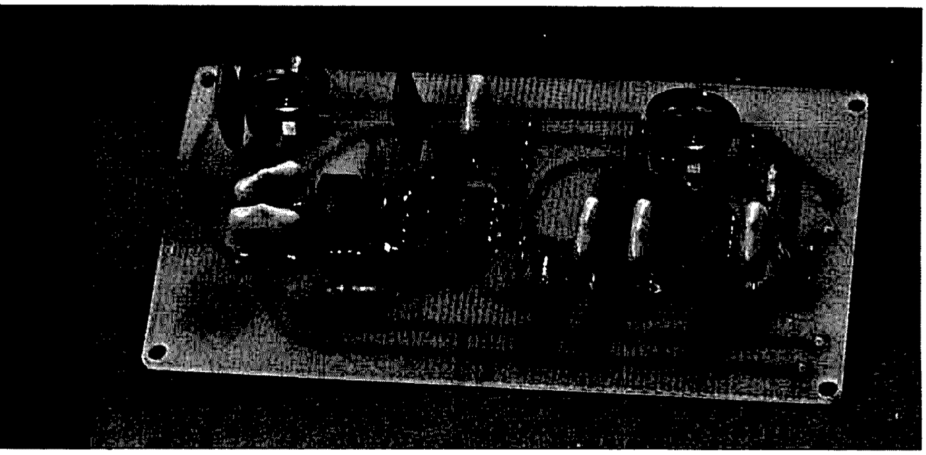
This CMOS crystal -frequency marker standard is a big improvement over the standard once used and consumes only 2.8 milliamperes at 9 volts, or 25.2 milliwatts. A 9-volt transistor radio battery power source was used and will probably last close to its shelf life. The frequency standard works fine through 10 meters. However, no equipment was available to check it through uhf, so I don't know whether sufficient energy will be present at these frequencies. The switching speed appears to be adequate, however.

Incidentally, a DT-cut crystal should be used for minimum temperature effects and frequency drift problems, since this type cut has much less frequency variation due to nearby heat sources.

## references

1. *The Solid-State Databook*, Series SSD203B, RCA Communications, Inc., pages 352-353.
2. Gerald Hall, K1PLP, "A TTL Crystal Oscillator," *QST*, February, 1974, page 34.

**ham radio**



# audio-frequency speech processor

Design and construction of a logarithmic limiting type speech processor featuring $\mu$A741 ICs

**An easy way to improve** ssb-transmitter performance is to use a well-designed speech processor between microphone and transmitter. Such a processor, if properly adjusted, will make a noticeable improvement in your transmitter speech readability. Contrary to current rumors, poor audio quality and increased bandwidth are not inherent in a good design.

Audio-signal conditioning requires no transmitter modifications with the exception of a cooling fan to protect components because of the increase in average power input to the final amplifier. Such a fan may be mounted externally if needed.

The speech processor described here is of the logarithmic type[1] and is easy to adjust and use. When used with my SB-401, this speech processor resulted in a signal-strength increase at the receiving end of at least two S-units. I've yet to get a report of poor audio quality or splatter, even from operators using monitor scopes. As a result, I use the processor at all times.

## response

Much has appeared in the amateur literature regarding the ideal response of speech processors. For example, reference 2 states that, "The only way to accurately evaluate the actual improvement of-

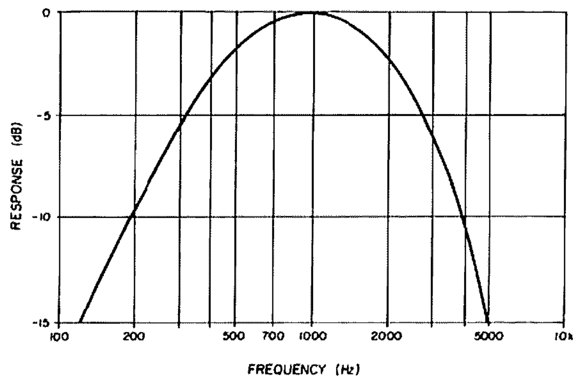



fig. 1. Speech-processor response below clipping level. $V_{in} = 0.054$ volt p-p; $V_{out} = 0.230$ volt p-p at 0 dB; Generator $Z_{out} = 50$ ohms; $Z_{load} = 1$ megohm.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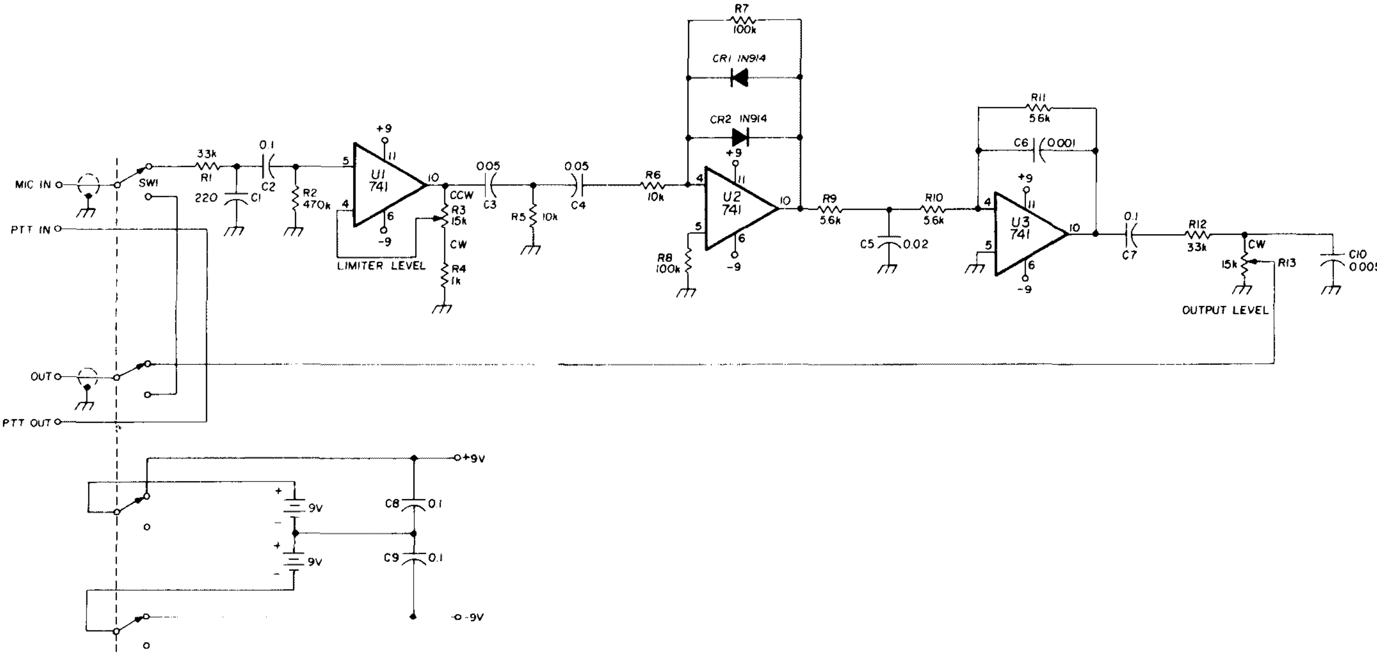



fig. 3. Speech-processor schematic. Numbering for 14-pin DIPs is shown.

| | |
|---|---|
| U1,U2,U3 | μA741 operational amplifier (use mini-dip for PC board) |
| SW1 | 4 pdt rotary switch |
| C1 | 220 pF, 50 WVdc disc ceramic |
| C2,C7,C8,C9 | 0.1 μf, 50 WVdc mylar |
| C3,C4 | 0.05 μf, 50 WVdc , 5% mylar |
| C5 | 0.02 μf, 50 WVdc, 5% mylar |
| C6 | 0.001 μf, 50 WVdc, 5% mylar |
| C10 | 0.005 μf, 50 WVdc mylar |
| CR1,CR2 | 1N914 diodes |
| R3,R4 | 15k or 10k linear pot, composition type |
| All other resistors | 1/4 watt, 5% |
| SW1 | 4 pole, 2 position |

fered by a speech processor is to measure the Intelligibility Threshold Improvement (ITI)." After much experimentation, I found the frequency response of **fig. 1** to be a good compromise between maximum readibility and good audio quality. The audio band width is sufficiently limited without impairing voice quality.

## circuit description

**Fig. 2** shows the block diagram for the processor. The schematic is shown in **fig. 3**. R1C1 forms an rf filter to prevent rf on the microphone cable from getting into the processor. C2R2 gives some highpass filtering action. R2 provides bias to the noninverting input of U1, and combined with the input impedance of U1, will provide sufficient load to a crystal or ce ramic microphone to attenuate low-frequency components and reduce harmonic distortion produced by very-low-frequency energy entering the microphone.

U1 is an adjustable gain-preamplifier, which overcomes losses in the highpass filter (C3, C4, and R5) and sets the input level to U2, the limiting amplifier. U2 limits because the nonlinear resistance characteristics of CR1 and CR2 supply increasingly heavier negative feedback as U2 output amplitude increases, which provides a logarithmic response. Limiting is soft and is similar to that of a compressor followed by a hard limiter. R8 eliminates any dc offset in U2 output due to input bias current. R9, R10, R11, and C5, C6, and U3 form a lowpass active filter, which attenuates frequencies above 2.8 kHz that may be generated in the clipping process. R13 provides output level adjustment, so that the transmitter drive-level control doesn't have to be changed when the processor is switched in or out.

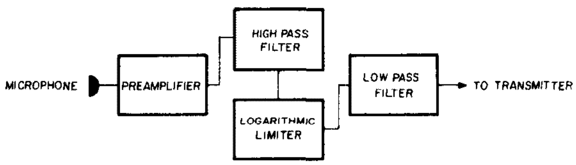



fig. 2. Block diagram of the logarithmic speech processor.

## construction

Wiring and component values with the possible exception of the resistors and capacitors that comprise the high- and lowpass filters are not particularly critical. The usual precautions applicable to audio circuitry should be observed.

The original version was wired on a small PC board intended for breadboarding DIP integrated circuits. It was then mounted along with two standard alkaline transistor radio batteries, in a 3 x 4 x 5-inch (77x102x 128mm) aluminum box with plenty of room to spare.

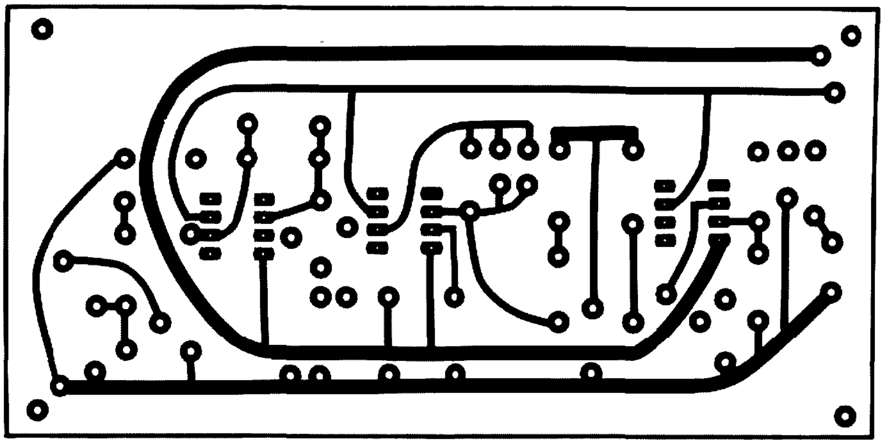


fig. 4. PC board layout, foil side.

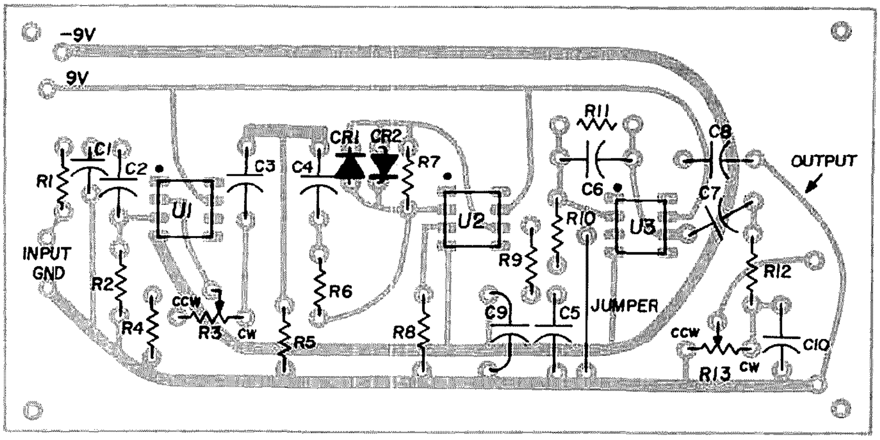


fig. 5. Component side of the PC board showing parts placement.

A PC board layout is shown in **figs. 4** and **5.*** I used IC sockets in my original unit, but these are not absolutely necessary.

The alkaline batteries should be good for about a year of normal operation, but an ac power supply could be substituted if precautions are taken to minimize ac hum, which would be particularly troublesome if introduced at this level.

I mounted two double-pin audio connectors of the same type used for the transmitter microphone connector on the rear of my processor. These are the input and output connectors and permit easy insertion and removal of the processor. The cable between processor and transmitter is of the same type as used on the microphone and is terminated at each end with a connector identical to that used on the microphone. (Don't forget to carry any PTT lines directly through the unit.)

*A printed-circuit board is available from the author for $4.95 and a self-addressed stamped envelope.

Although holders are available for 9-volt batteries, I found that the heavy flat plastic wire ties used by electricians to bundle conductors make a very secure mounting. Make sure to use the type with a mounting hole molded in one end. Standard battery clips were used for the electrical connections.

## operation

Operation is simple. Once the settings of R3 and R13 have been established, the on-off switch is the only control used. I used screwdriver-adjust pots in my unit.

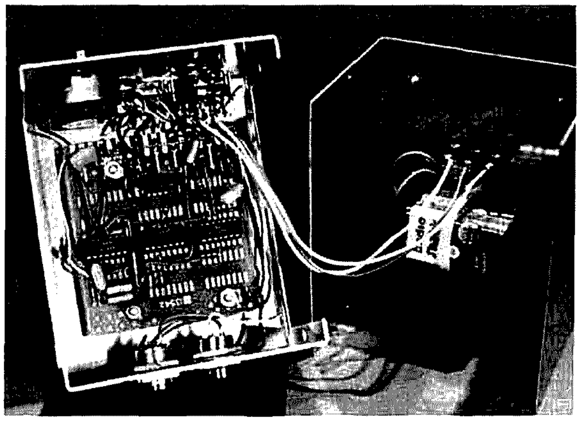
The transmitter should be tuned in the normal fashion with the processor switched out. The transmitter microphone gain control should be set at the proper level, using a monitor scope or the transmitter's ALC indicator according to the instruction manual. Turn limiting control R3 about one-half turn clockwise, switch in the processor, and adjust output control R13 for the same ALC indication or peak-monitor-scope deflection that was obtained

without the processor. On-the-air reports can then be used to determine the optimum setting of R3.

After adjusting R3, always recheck the setting of R13. Use discretion in adjusting R3. Extreme limiting should be avoided if the best combination of signal strength and audio quality is to be attained.

A too-high setting of R3 will also cause normal background noise to become objectionable as the audio gain for low-level signals is increased. Consequently, the processor can be used most effectively in a quiet, echo-free environment.

The use of this speech processor will increase the average power input to the final amplifier and cause some increase in final-amplifier -tube dissipation. To extend tube life, I added a small cooling fan. The fan I used is of the axial type. It is rectangular, about 5 inches (128mm) on a side and about 1½ inches (38.25mm) thick, with mounting holes in each corner. I fastened four small rubber feet at the mounting holes and placed the fan on top of the perforated lid of the transmitter, directly above the final tubes, so that the fan draws air from inside the cabinet. I decreased the air flow and noise level of the fan, both of which were excessive, by installing a resistor of about 375 ohms in series with the fan motor, which resulted in a quiet cooling system.



**Breakaway section showing wiring and underchassis layout.**

Performance over a 12-month period has been excellent. The increased range and number of solid ssb contacts have more than justified the modest investment in time and material for the speech processor.

## references


1. Bob Myers, W1FBY, "A Quasi-Logarithmic Analog Amplitude Limiter With Frequency-Domain Processing," *QST*, August, 1974, page 22.
2. Gene Nurkka, VK9GN, "Integrated-Circuit Single-Sideband Speech Processor," *ham radio*, December, 1971, page 31.


**ham radio**

# low-cost microwave spectrum analyzer

How to put together a microwave spectrum analyzer from surplus odds and ends — the completed unit covers dc to 2 GHz

**My ongoing efforts to develop** low-cost modules for the 1296-MHz band have often required the use of a spectrum analyzer for monitoring (and minimizing) harmonic and spurious frequency components. Numerous excursions through the local surplus test equipment emporiums revealed that an acceptable instrument could cost several thousand dollars — well beyond the budget of the most dedicated experimenter. While searching for the unbeatable surplus buy which never materialized, I noticed the ready availability and comparatively low cost of a wide variety of S-band (2-4 GHz) test instruments and components. It occurred to me that a microwave spectrum analyzer could be put together from these available parts at a considerable savings. This article documents the design, construction, operation, and performance limitations of the resulting microwave spectrum analyzer. While I doubt that any reader will want to duplicate my design in its entirety, I hope this article will provide guidance and encouragement to anyone attempting a similar project.

## performance requirements

The operation of any spectrum analyzer can be characterized in terms of its frequency coverage, dispersion, dynamic range, sensitivity, and resolution. To display frequency components well into the microwave region, I designed my spectrum analyzer to cover dc to at least 2 GHz. The same design strategy could be easily applied to other frequency bands. In fact, the upper frequency limit of this analyzer was later extended to 2.5 GHz, as discussed later.

Dispersion describes the ability of a spectrum analyzer to display a broad slice of the frequency spectrum in a single sweep. Many of the low-cost analyzers on the surplus market display only a few MHz at a time. Such narrow-dispersion spectrum analyzers are useful as panadaptors, which display all signals within several hundred kHz of a specified operating frequency, but when tuning a microwave local-oscillator chain, monitoring mixer image response, or measuring transmitter harmonic content, it is often desirable to display a band several hundred (or even thousand) MHz wide. The spectrum analyzer shown here can display the spectrum from dc to 2 GHz in a single sweep. Since it's often desirable to narrow this sweep for a closer look at a particular signal, variable dispersion capabilities are included in the design.

Sensitivity and dynamic range define the minimum and maximum signal amplitudes which an analyzer can display without distortion. In accordance with good engineering practice, I try to suppress all transmitted spurious products by 50 dB or more. To accurately measure this performance, the spectrum analyzer requires at least 50 dB of dynamic range. As

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for maximum input level, I often want to display a +10 dBm (10 mW) signal (this is the local-oscillator injection level required of many balanced mixers). Thus, 50 dB dynamic range with a +10 dBm maximum input level yields an ultimate sensitivity requirement of −40 dBm, or 0.1 μW. Greater sensitivity (a lower minimum discernible signal) could have been obtained, but only at the expense of dynamic range.

The objective of any spectrum analyzer is to display the various components of a complex waveform in the frequency domain. The closer the frequencies of any two components, the more difficult it is to separate them on the spectral display. Resolution relates to the minimum frequency separation between two signals of equal amplitude which will still permit the operator to discern two separate frequency components on the display.

Resolution can be approximated as twice the i-f bandwidth of the analyzer system. Generally, the objectives of wide dispersion and narrow resolution are mutually exclusive. When measuring transmitter audio intermodulation products with a two-tone test, for example, a resolution of a few hundred Hz is required, and dispersion is likely to be several tens of kHz. When viewing harmonics of a 100-MHz oscillator, on the other hand, 2-GHz dispersion may be required, but a resolution of several tens of MHz is acceptable.

Resolution is primarily a function of i-f bandwidth, which for my analyzer is fixed at 1 MHz. Therefore, with this design, I can resolve frequency components to within about 2 MHz.

As most amateurs know, a general-coverage communications receiver can be used as a rudimentary high-frequency spectrum analyzer. With an input signal applied to the antenna terminals, the receiver is manually tuned through its frequency range (dispersion). Frequency components are detected and displayed (perhaps with the receiver's S-meter). Resolution is a function of the i-f bandwidth, which is probably a few kHz. Sensitivity is a function of the receiver's noise floor, and dynamic range is limited by the receiver's agc and overload characteristics. Obviously, wide dispersion measurements require considerable operator intervention, in the form of tuning. Gain variations of the receiver from band to band will limit the accuracy of its amplitude indication. Additionally, any nonlinearity in the receiver's agc circuit may prevent accurate amplitude measurement across the receiver's entire dynamic range. Also, the receiver's image and spurious rejection may be insufficient to eliminate false indications.

Ideally, a workable spectrum analyzer should be a superheterodyne receiver in which these shortcomings are minimized. Frequency tuning should be both automatic and rapid. Instead of an S-meter, amplitude is displayed on an oscilloscope. If the scope's horizontal deflection is slaved to the receiver's tuning mechanism, the result is a display in the frequency domain. Dynamic range must be max-

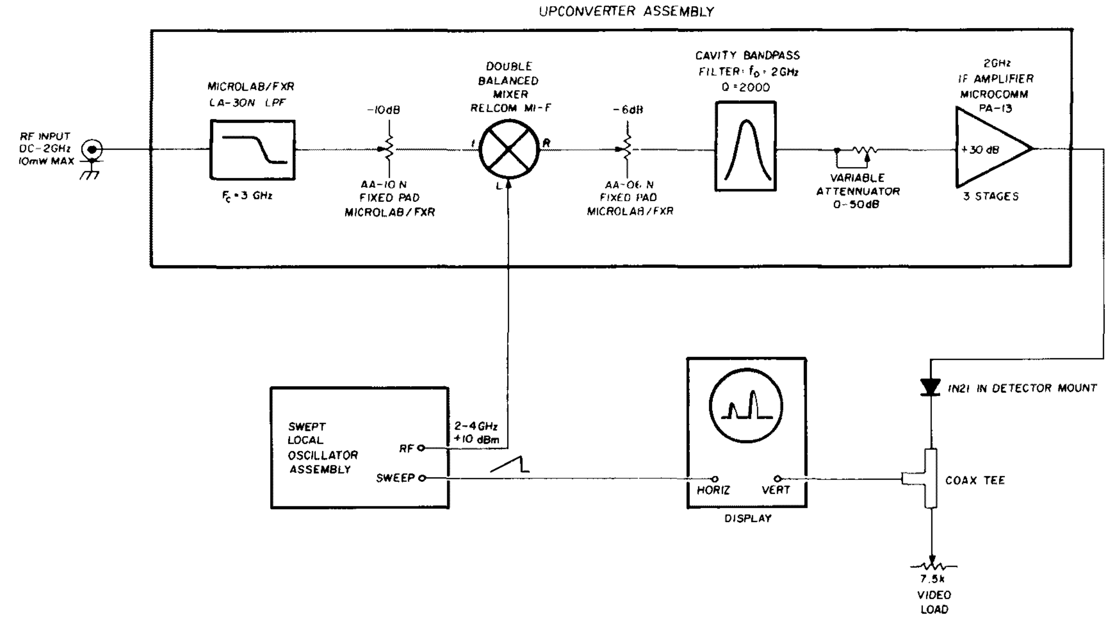
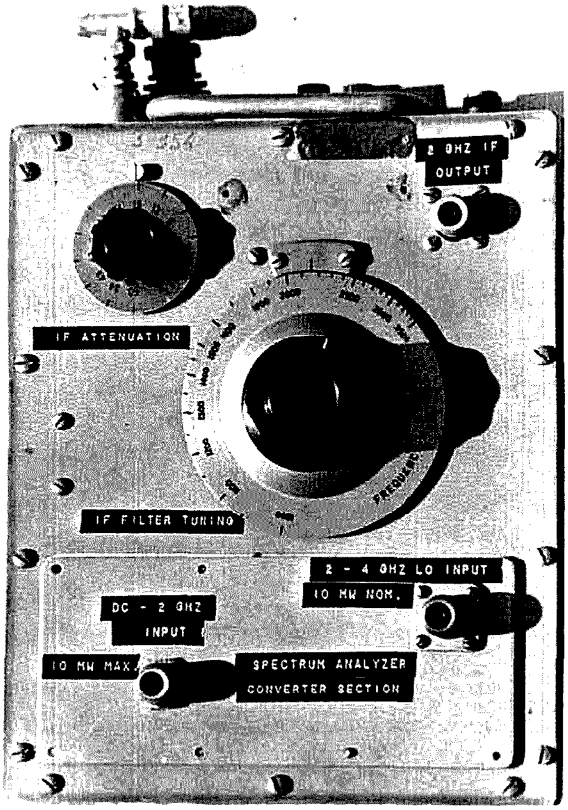



**fig. 1. Block diagram of the microwave spectrum analyzer. Most components were purchased on the surplus market. The swept local oscillator, the key to the analyzer, is a surplus 2-4 GHz backward-wave oscillator. The display is an ordinary oscilloscope with dc coupling and provisions for external sweep.**




Front panel of the microwave spectrum analyzer showing the operating controls. The variable i-f attenuator (upper left), although calibrated in 6 dB steps, permits amplitude comparison of signals displayed on the oscilloscope. The i-f bandpass filter (center) sets the frequency coverage of the analyzer, as discussed in the text.

imized, and spurious/image responses eliminated, to the greatest possible extent.

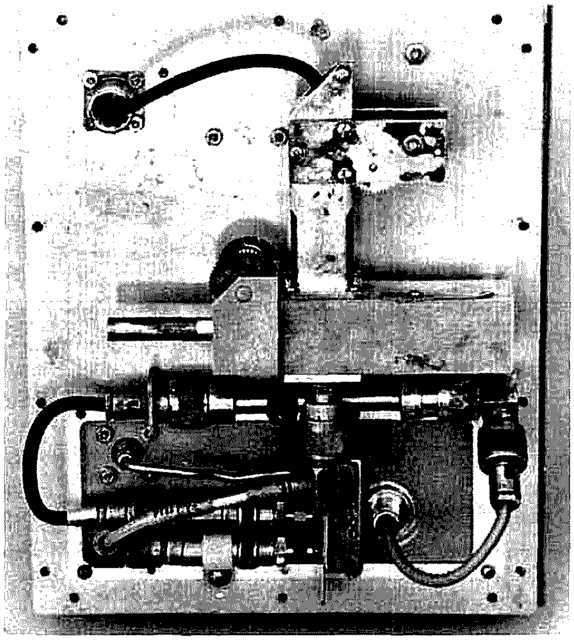
Many of the objectives discussed previously are met in the design shown in **fig. 1**, a wide dynamic range microwave receiver with an electronically tuned local oscillator and ample image rejection. It includes a 2-GHz i-f amplifier with variable gain and fixed bandwidth, and a sensitive detector for driving an oscilloscope. Unlike conventional receivers, this design up-converts the incoming signal to an i-f in the microwave region. Although this approach complicates i-f design, it permits wide dispersion tuning. It also improves separation of the rf and image signals so a simple lowpass filter can be used to eliminate image responses.

The microwave spectrum analyzer is divided into three separate sub-systems: the local oscillator, unity-gain upconverter, and display sections.

## local oscillator

Central to the design of this spectrum analyzer was the availability, on the surplus market, of a leveled, swept signal source covering 2 to 4 GHz. I used an Alfred 622BK sweep oscillator, but any similar generator should work satisfactorily. These sweep generators consist of a voltage-controlled oscillator, typically a backward wave oscillator or BWO (a microwave oscillator built around a device similar to a traveling wave tube), power supplies, a sawtooth generator for developing a constantly varying vco control voltage, and leveling circuitry to maintain constant output across the band. *Start* and *stop* frequency adjustments permit the oscillator to sweep all, or any portion of, the 2 to 4 GHz band. Leveled output power is typically 10 to 30 milliwatts.

Many companies are currently retiring their BWO sweep generators in favor of wideband, solid-state units, so quite a few BWO generators have recently appeared on the surplus market at prices ranging from $200 to $400 or so. Since this is the most costly component of the microwave spectrum analyzer, make sure the unit you buy is in good operating condition. Reputable electronics surplus dealers will often let you power up an instrument and make a few measurements prior to purchase. A practical test requires the use of a microwave power meter (bolometer bridge or equivalent) to observe output power in the leveled CW mode as the generator is manually tuned across the band. Although a few dB variation is acceptable, dead spots or severe power drop-off at the high end of the band indicates a failing BWO. A good, used BWO should provide years of reliable life in intermittent amateur service.



Interior of the spectrum analyzer converter section showing the double-balanced mixer and attenuation pads, input filter, i-f filter, and variable i-f attenuator.

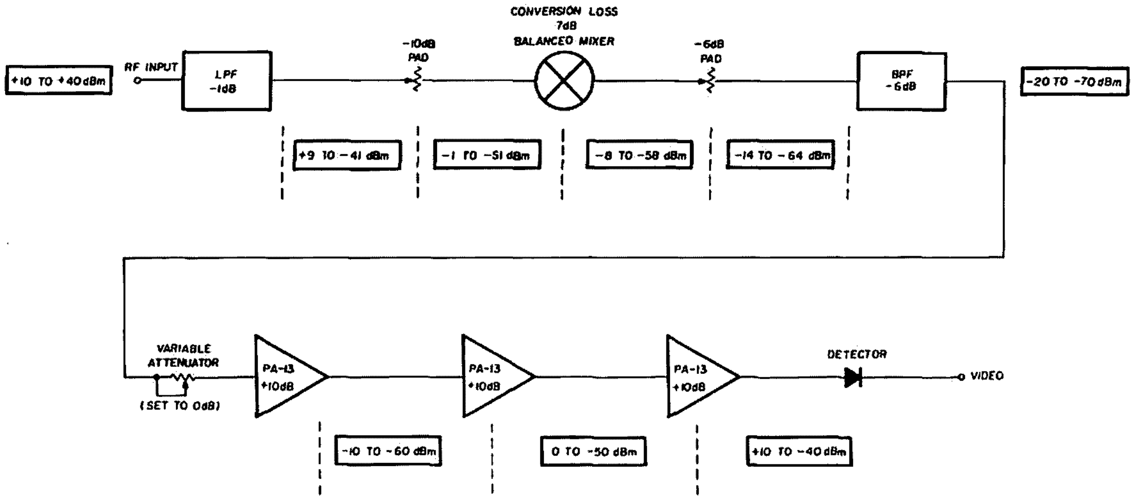


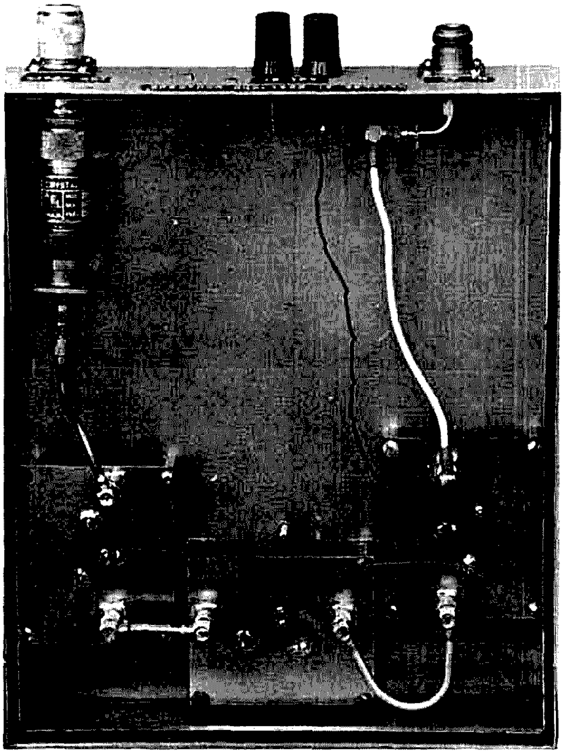

**fig. 2. Power levels in the upconverter used with the microwave spectrum analyzer (with the variable attenuator set at 0 dB, which results in unity conversion gain).**

As can be seen in the block diagram, **fig. 1**, the spectrum analyzer converter assembly consists of an input lowpass filter, an S-band double-balanced mixer, some attenuator pads, a high-Q i-f filter, a variable i-f attenuator, and sufficient i-f amplification to bring maximum conversion gain to unity. The i-f detector and its video load, though installed in the converter assembly, are discussed later with the display.

The characteristics of the balanced mixer, more than any other component, establish the linearity and dynamic range of the analyzer. I used a Relcom M1F mixer which I found on the surplus market for $35 (the mixer retails for about $200). The rated frequency response of this mixer is dc-2 GHz at the i-f port, and 2-4 GHz at the rf and LO ports. Note that the incoming signal is applied to the *i-f port*; the *rf port* drives the i-f system. Thus, all ports are operated withinn their specified frequency ranges.

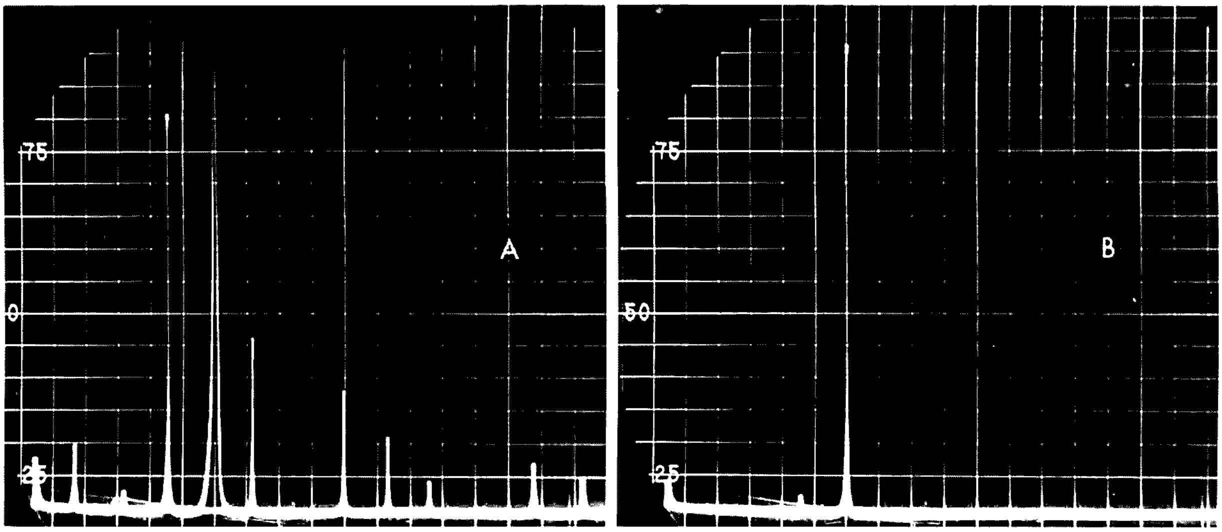
With the 10 mW of local-oscillator injection applied to the mixer from the sweep generator, the mixer's conversion efficiency is compressed by 1 dB at an input signal level of 1 mW. Since I wanted to analyze a 10 mW signal on the spectrum analyzer without exceeding 1 dB compression, it was necessary to place a 10 dB attenuation pad ahead of the mixer's input (i-f port). This pad also assures proper impedance termination for the mixer, as does the 6 dB attenuator at the output (rf port). Fixed attenuators for dc to 2 GHz are available to the surplus bargain hunter for as little as $5.00, or may be purchased new for $15 to $20.

With 2 GHz i-f, and a swept LO covering 2 to 4 GHz, the mixer will respond to signals in the dc-2 GHz region, as well as in the 4-6 GHz image band. Any components in the image band will cause considerable confusion on the display. Thus an input lowpass filter was installed in the system to block all signals above 3 GHz from entering the mixer. I used a Microlab/FXR LA-30N filter which I salvaged from another piece of equipment. Although the filter originally cost $40, similar devices are available through surplus outlets for $5 to $10.



**Spectrum analyzer i-f section. The three 2-GHz amplifiers are at the bottom. Diode detector is at upper left.**

The i-f bandwidth of this analyzer is established by a high-Q tunable coaxial or cavity filter which is tuned to 2 GHz. I used the filter from a surplus TS-406 noise generator, but any cavity with a Q of 1000 or greater should be acceptable. It's also possible to use filter to vary i-f gain . The attenuator I used was also salvaged from the TS-406 noise generator, but any continuously variable or step attenuator rated to 2 GHz is acceptable — 10 dB steps will allow coarse system gain control; if 1 dB resolution is included,

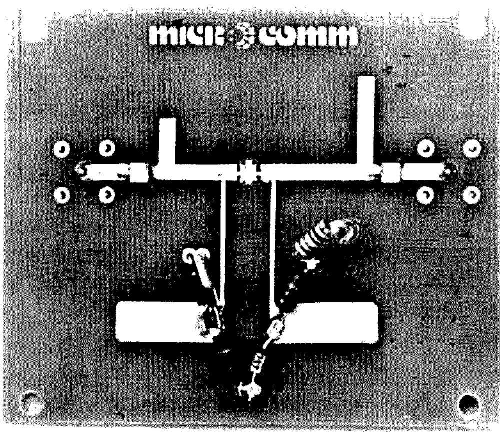



**Example of spectral impurity, as displayed on the microwave spectrum analyzer. Presence of harmonic, subharmonic, and spurious signals shown in *A* is the result of an overdriven uhf amplifier. The same amplifier, with drive reduced to the rated level, is shown at *B*; the one spurious component is down by more than 20 dB. Display is from dc to 2 GHz.**

a *transmission-mode* cavity wavemeter as an i-f filter. These widely available devices have a loaded Q of several thousand, and exhibit only a few dB of insertion loss at resonance. Note that an absorption-type wavemeter is *not acceptable* because the filter must pass maximum signal to the i-f amplifiers at resonance.

A 50 dB variable attenuator was installed after the the attenuator can also be used for accurate signal level comparison. This is accomplished by viewing one signal component, setting a convenient reference level on the display, varying the attenuator for a like indication on the other signal component, and noting the change in attenuator settings.

Considerable i-f gain is required to achieve the desired sensitivity. I cascaded three stages of the Microcomm PA-13 buffer amplifier.* These microstripline amplifier modules offer 10 dB of gain per stage across the 2.0-2.3 GHz band, and are biased for 30 mW output at 1 dB gain compression. Since i-f noise figure is not a limiting factor so far as system sensitivity is concerned, any available wide dynamic range amplifier for 2 GHz may be used.

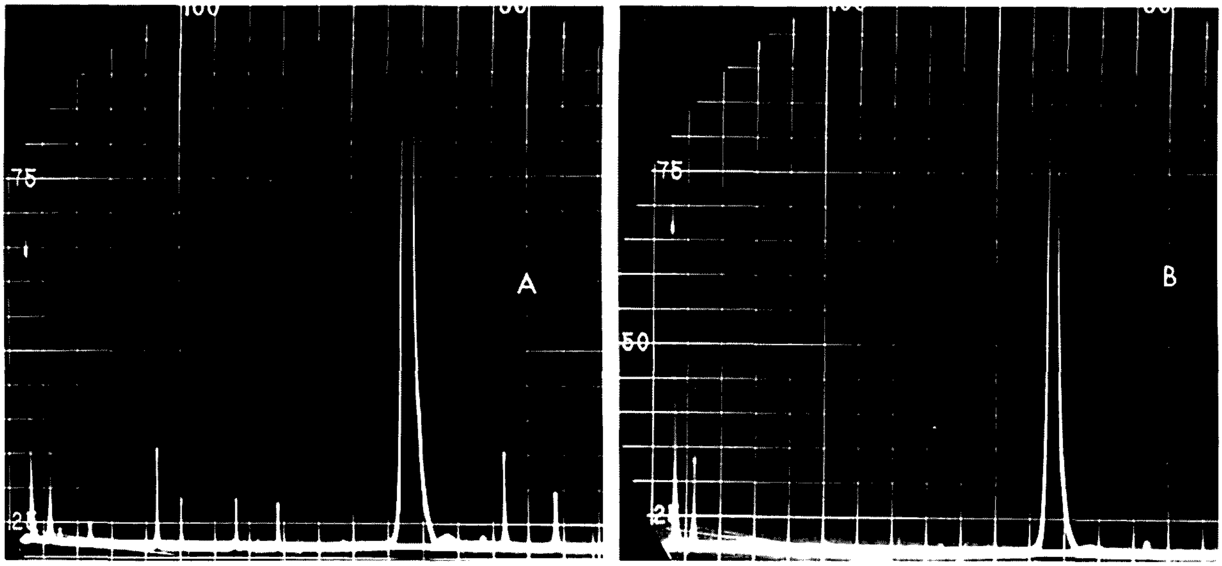


**Microstripline side of the 2-GHz amplifier (Microcomm PA-13). Three of these units provide 30 dB gain at 2 GHz.**

## display

The local-oscillator signal for the spectrum analyzer is swept by a sawtooth waveform; therefore, displaying a signal in the frequency domain is simply a matter of detecting the output signal from the i-f amplifiers, applying the recovered video to the vertical deflection amplifier of an oscilloscope, and applying the sawtooth output voltage from the sweep generator to the oscilloscope's horizontal axis. Since a relatively slow sweep rate is used, the

*Available for $64.95 per stage (plus postage and handling) from Microcomm, 14908 Sandy Lane, San Jose, California 95124.




Output of a local-oscillator chain for a 1296-MHz converter. At *A* the i-f gain of the spectrum analyzer has been increased to show the spurious signals, which are 30 dB below the desired signal. The display at *B* shows the output of the same local oscillator after it has passed through a 3-pole bandpass filter. Although the filter has attenuated the desired signal by 1 dB, all spurious signals are down more than 50 dB.

frequency response of the oscilloscope is unimportant. Virtually any scope with dc coupling and provisions for external sweep may be used.

The dynamic range of the detector which follows the i-f amplifiers is of major concern. **Fig. 2** shows the nominal gain or loss of each element of the analyzer upconverter, as well as maximum and minimum signal levels present at each stage with zero i-f attenuation (maximum sensitivity). Since the upconverter is operated at unity gain, the power available to the detector will vary from +10 to −40 dBm. Thus the diode's tangential sensitivity must be considerably below −40 dBm, and the diode's saturation point above +10 dBm, for a usable display. Although I know of no diode whose transfer characteristics are uniform over so wide a range, the 1N21 family of point-contact diodes are acceptable within certain limitations (discussed later).

Diode dynamic range is enhanced by the optimum terminating impedance, which may vary between 1000 ohms and 10 kilohms or so. The input impedance of an oscilloscope's vertical deflection amplifier is typically 1 megohm; thus, to assure proper termination for the diode, a loading resistor is required, as shown in **fig. 1**. Since terminating the diode's video port degrades the amplitude of the recovered video, increased oscilloscope vertical sensitivity is required. On my analyzer, a 7.5k ohm video load, in conjunction with a vertical sensitivity of 10 mV/cm, provides an acceptable display.

Note that the vertical display of the spectrum analyzer is approximately linear, not logarithmic. Therefore, it is possible to view only about 25 dB of amplitude range at once, and only with extremely limited amplitude resolution. However, by using the i-f attenuator to establish reference levels as described previously, the entire 50 dB of usable

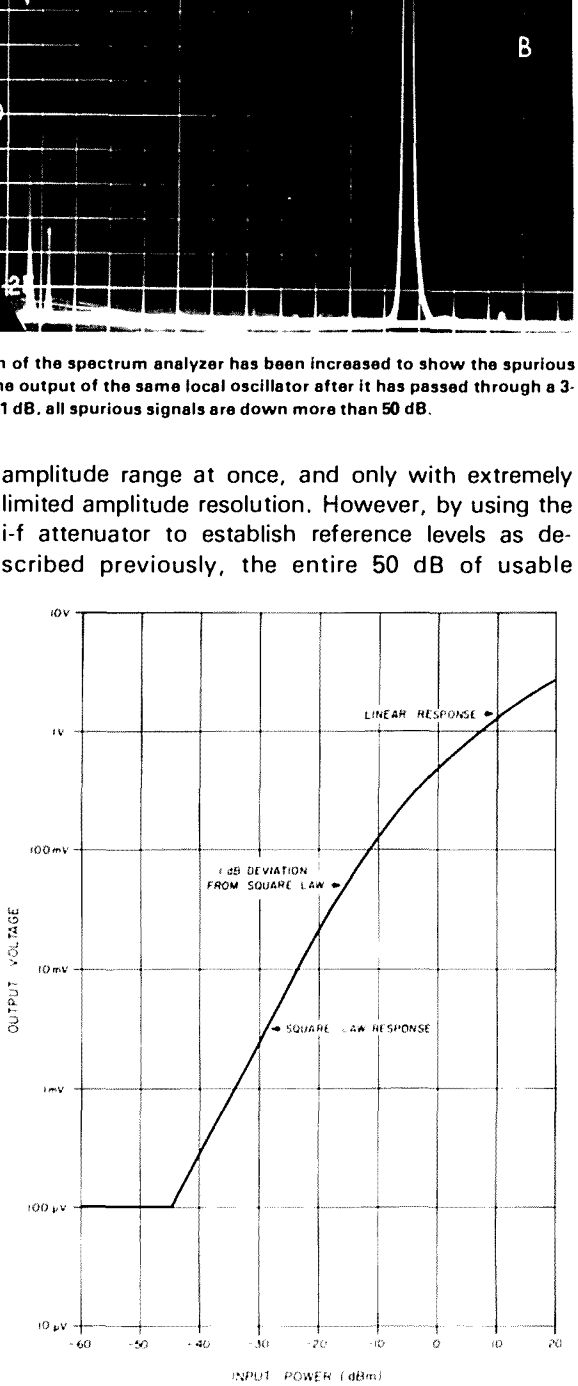



**fig. 3. Transfer characteristics of a typical microwave diode detector using a point-contact diode.**

dynamic range can be put to use. Future plans include capability for a logarithmic display, as outlined toward the end of this article.

I built the entire upconverter module in the case of a surplus noise generator (the filter and attenuator of which formed key i-f elements in my system). As can be seen in the photographs, the i-f amplifiers are mounted on standoffs inside the main chassis, and connected with UT-141 semi-rigid coaxial cable. If flexible coax is used, I recommend RG-142B/U. This ¼-inch (6.5mm) cable is double-shielded, silver plated, has a Teflon dielectric, and accepts clamp-type SMA plugs of the low-cost E. F. Johnson JCM series.

If desired, the i-f amplifiers may be mounted in Pomona 3601 die-cast aluminum boxes. These boxes present a neat appearance and afford somewhat better shielding than the standoff approach I used.

The various components of the upconverter assembly sport a variety of connectors; between-series adapters are required to interface types N, SMA, and TNC receptacles.

## operation and applications

A recent article discussed a variety of spectrum analyzer applications of interest to amateurs.[1] For additional information, Hewlett-Packard has published two application notes which discuss both the procedures and the theory of spectrum analysis.[2,3]

The accompanying photographs show the displays obtained with this analyzer when measuring LO harmonic content, balanced mixer carrier rejection, converter image response, and transmitter intermodulation products.

## improving sensitivity

One of my primary design objectives was the ability of the analyzer to handle relatively large (+10 dBm) input signals without overloading. The input pad shown in **fig. 1**, although it prevents mixer overload at these signal levels, obviously limits the ultimate sensitivity of the system. Additional i-f amplification would enhance the ability of the analyzer to display very small signals — but then the detector diode would saturate at high input signals.

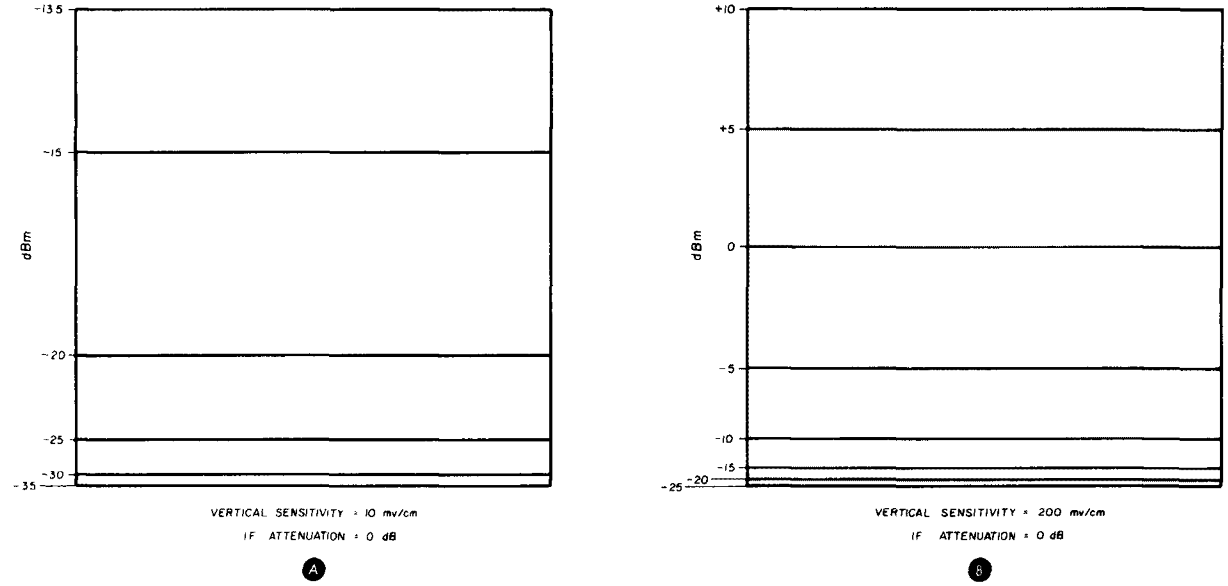
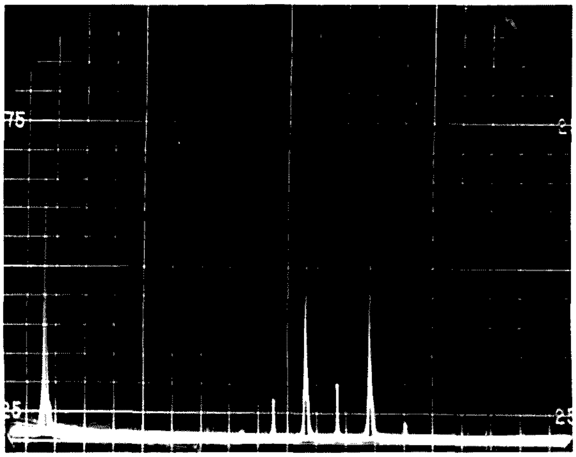



**fig. 4. Approximate vertical scale calibration of the analyzer display (no video processing). Note the excellent linearity at high power levels, and the compressed display when the detector diode is operated in the square-law region.**

Where both increased sensitivity and large-signal handling capability are required, it's necessary to replace the 10 dB input pad with an appropriate step attenuator (and possibly adding additional i-f stages). One candidate for the input attenuator is the Kay 520 which offers 10 dB steps to 70 dB and is flat to 2 GHz. This unit is priced under $100, but like everything else in my spectrum analyzer, units are often available through surplus sources at considerable savings.

Note that no variation in input attenuation or i-f gain can increase the dynamic range of this analyzer beyond 50 dB or so. Therefore, it is essential that the operator select input attenuation which is appropriate for the anticipated signal level. In short, input attenuation should be such that the signal applied to the mixer does not exceed 1 mW, and that to the detector remains below 30 mW. A simple operating check involves increasing the input at-



As a transmitting mixer's i-f port is overdriven, intermodulation products at *LO* ± *2i-f* become more pronounced, and the amplitude of signals at *LO* ± *3i-f* begins to increase. This display also shows the second harmonic of the i-f injection signal (at the left).

tenuation by 10 dB while decreasing i-f attenuation by the same amount. An increase in the apparent amplitude of the displayed signal indicates that the mixer was being over-driven.

## expanding frequency coverage

Recently I became involved in designing amplifier, mixer and LO modules for 2304 MHz, and wanted to extend the upper frequency limit of this spectrum analyzer. This could be accomplished by varying either the swept LO frequency or the i-f frequency, or both. Since the LO frequency range is limited by the coverage of the available sweep generator, I chose to change the i-f frequency.

With a 2 to 4 GHz swept LO, frequency coverage from 500 MHz to 2.5 GHz could be obtained by modifying the analyzer's i-f to 1.5 GHz. However, this exceeds the rated frequency range of the mixer's rf and i-f ports. Fortunately, the frequency response of the i-f port of the mixer I used exceeded the specified 2 GHz. At 2.5 GHz input, vswr is degraded somewhat, but the use of the 10 dB input pad effectively masks this mismatch. As for the frequency response of the mixer's rf port (used to develop i-f output), reducing the i-f frequency to 1.5 GHz degrades conversion efficiency by several dB. The i-f is fixed, however, so this degradation applies equally to all input signals, and no system linearity is sacrificed.

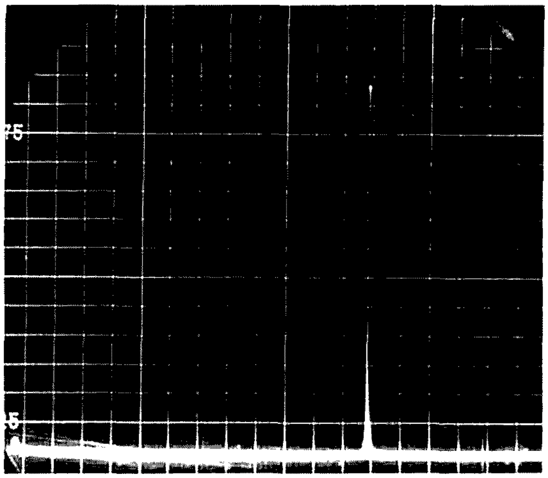
I originally planned to switch in a separate i-f system for high band (0.5-2.5 GHz) coverage, but I discovered that the PA-13 i-f amplifiers were sufficiently broadband that they have usable gain at 1.5 GHz. Since the cavity filter used to establish i-f bandwidth is tunable, setting the spectrum analyzer up for high-band coverage is simply a matter of retuning the i-f filter to 1.5 GHz. In this mode, overall analyzer sensitivity is degraded by about 10 dB, but it is still adequate for many applications.

With a 1.5 GHz i-f, the input signal range extends from 500 MHz to 2.5 GHz, and the image band from 3.5 to 5.5 GHz; therefore, the existing 3-GHz lowpass filter provides ample image rejection without degrading input frequency coverage.

## detector limitations

In normal operation (**fig. 2**) the diode detector sees a power level from −40 to +10 dBm. **Fig. 3** shows the transfer characteristics of a typical microwave detector diode over this range. It can be seen that the diode is being driven from its square-law region into its linear region. Thus, at high signal levels, 10 dB of signal change results in approximately a 10 dB change in recovered video amplitude; at lower power levels a 10 dB input signal variation may change video amplitude by up to 20 dB. Obviously, the linearity of the display is marginal, at best, and usable dynamic range is restricted to about 25 dB unless the operator varies i-f gain, or vertical sensitivity or both.

Hewlett-Packard introduced a series of oscilloscope overlays for interpreting non-linear



Properly driven 1296-MHz transmit mixer, at the output of a 3-pole bandpass filter. The image, i-f, and LO signals are all 40 dB down; intermodulation products are more than 50 dB down. The small pip at the left side of the trace is the bandedge marker and represents zero frequency; it is produced when the LO sweeps through the i-f filter.

(more properly, non-logarithmic) swept displays.[4] A similar set of overlays, which I have derived for my spectrum analyzer, is shown in **fig. 4**. Although this calibration data is valid only for my analyzer, a similar vertical axis can be derived for any spectrum display. All you have to do is apply an input signal of known amplitude through an accurate step attenuator. By varying the attenuation and noting the

displayed amplitude, calibration lines can be grease-penciled directly on to the face of the CRT.

The utility of this spectrum analyzer would be greatly enhanced if it were possible to display simultaneously all signals between −40 and +10 dBm. If you want to view the entire 50 dB dynamic range without adjusting reference levels with the i-f attenuator or varying vertical sensitivity, it will be necessary to apply the output of the detector into a compression video amplifier.

There are several integrated circuits available which provide logarithmic video amplification; with a logarithmic amplifier a display such as that shown in **fig. 5** can be obtained. Note that below about −10 dBm, the display approaches a uniform 5 dB per centimeter deflection. However, the transition from square-law to linear detection results in severe scale compression at higher power levels.

An ideal spectrum analyzer display should have vertical response similar to that shown in **fig. 6.** Although I have not yet been able to achieve this performance, it should be possible by developing a logarithmic video amplifier which makes its transition to linear response above a selected input level.

An approach used successfully by Pacific Measurements in their logarithmic power meters involves an operational amplifier in which the feedback resistance is a nonlinear element (a semiconductor junction). As the junction potential of this feedback path is exceeded, the gain curve of the op amp changes. The result is an amplifier which makes its transition from logarithmic to linear response at selected power level. Perhaps some reader will be able to contribute a similar circuit for appropriately shaping the video output of the spectrum analyzer's diode detector. What is needed is a display whose

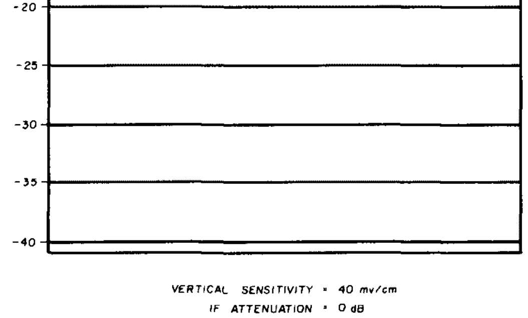



**fig. 5. Typical vertical scale calibration of a spectrum analyzer with a logarithmic video amplifier. In this case display linearity is good at lower input signal levels, but compresses rapidly as the detector diode enters the linear region.**

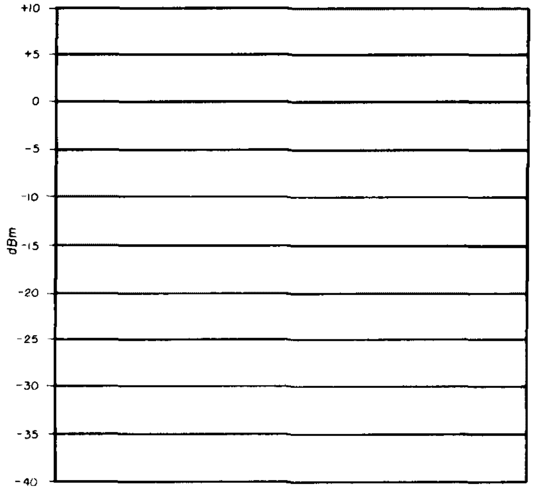



**fig. 6. Ideal vertical response for a spectrum analyzer. This response requires a video amplifier with a logarithmic compression curve below an input of 200 mV, and linear response above 400 mV.**

amplitude is graduated at 5 dB per centimeter, over the entire dynamic range of the system. This modification would significantly enhance both measurement accuracy and ease of operation.

## acknowledgements

Thanks are in order to Nick Marshall, W6OLO, for first encouraging me to try to build my own spectrum analyzer, and to Richard Chatelain, WB6JPY, who took the photographs. I must also acknowledge the eager support of my wife, WA6PLF. She reasoned that, if I built my own analyzer rather than spending funds I didn't have to buy one I couldn't afford, we would make good use of the money we saved. Although I'm not sure I understand the economics, I'm enjoying both the homebrew spectrum analyzer and our new car.

## references

1. Courtney Hall, WA5SNZ, "Understanding Spectrum Analyzers," *ham radio*, June, 1974, page 50.
2. *Spectrum Analysis*, Application Note 63, Hewlett-Packard, 1501 Page Mill Road, Palo Alto, California, May, 1965.
3. *More On Spectrum Analysis*, Application Note 63A, Hewlett-Packard, Palo Alto, California, November, 1965.
4. *Swept Frequency Techniques*, Application Note 65, Hewlett-Packard, Palo Alto, California, August, 1965.

**ham radio**

# serial converter

## for 8-level teleprinters

### This converter translates Baudot to ASCII and ASCII to Baudot using readily available ICs — recommended for the experienced amateur only

**At the outset I'd like to stress** that this is a project for the experienced amateur with the technical know-how to connect the converter described to appropriate points in his demodulator or fsk circuit.

The heart of the converter is the universal asynchronous receiver/transmitter (UAR/T), a 40-pin IC that contains both an independent, 8-bit asynchronous, digital-data receiver, and an 8-bit asynchronous, digital-data transmitter. The UAR/T has been described earlier in *ham radio*.[1,2]

Parts layout and wiring of the converter is quite a task in itself. Sockets are mandatory for the ICs as some are MOS devices, which are sensitive to ungrounded soldering irons. Fairly heavy bus wire is necessary for ground (common) and +5-volt leads. Also, liberal use of 0.01-$\mu$F ceramic disc capacitors (not shown in the schematic) is required to bypass +5, and −12 volt circuits.

All NAND gates used as inverters can be replaced by hex inverters to reduce component count. Type 74121s can be replaced by 74123s except where both A inputs are used.

Those readers interested only in receiving RTTY with an 8-level machine can save much time and money by deleting the connections (except for grounds and clock inputs) to the much more complex ASCII-to-Baudot section, which is shown below the dashed lines of both UAR/Ts in **fig. 1**.

The cost for the complete converter should be well below $100; possibly around $50. The 8223 proms can be replaced with 82S23s, which are currently on the market for about $3 each. The UAR/T ICs can be replaced with the GI AY-5-1013A, which is less expensive, about $6.50. The 3351 fifo devices can be obtained for about $14 through W6KS, as mentioned in the *RTTY Journal*. The other chips are standard devices and are available for about 50 cents or so.

Baudot-to-ASCII/ASCII-to-Baudot converters have been described in other publications but were not directly compatible with RTTY, which is a serial system, and didn't take advantage of the UAR/Ts.

### circuit description

The converter schematic is shown in **fig. 1**. The serial 5-bit Baudot signal at TTL level enters U1 at pin 20 and appears in parallel form at pins 8 through 12. ICs U2, U3 sense whether letters or figures have been sent and set R/S flip-flop U4 to enable either U5 or U6, which translate the 5-bit Baudot code to 7-bit ASCII code; this translation appears at U6 pins 1 through 7. (The no. 8 bit is a parity bit, which is used with computers, and is unnecessary for amateur RTTY. I have modified my model 35ASR so that the no. 8 bit is always a zero.)

Output from U6 is applied to UAR/T U7 and appears in serial form at pin 25, which is connected through driver Q1 to a 4N33 opto-isolator, U8. Another 4N33, U9, is connected in series with U8, whose output with that of U8 is inserted into the loop of a model 33 or 35 ASR. This circuit keeps modifications of the ASCII machine to a minimum.

The signal originating from either the keyboard or tape reader of the ASCI machine is connected to the input of the second 4N33, U9, and its output serially feeds into U7-20. This data appears in parallel form at U7 pins 12 through 19.

Because there are no *LTRS* or *FIGS* in the ASCII code, these characters must be generated. For this reason the 6th and 7th ASCII bit, which appear at pins 7 and 6 respectively to U7, are sensed by 7474 flip-flop U10. In combination with 7474 flip-flop U11, U10 will disable the 8223 proms U12, U13 temporarily.

**By Eric Kirchner, VE3CTP**, Ontario Science Center, Don Mills, Ontario, Canada

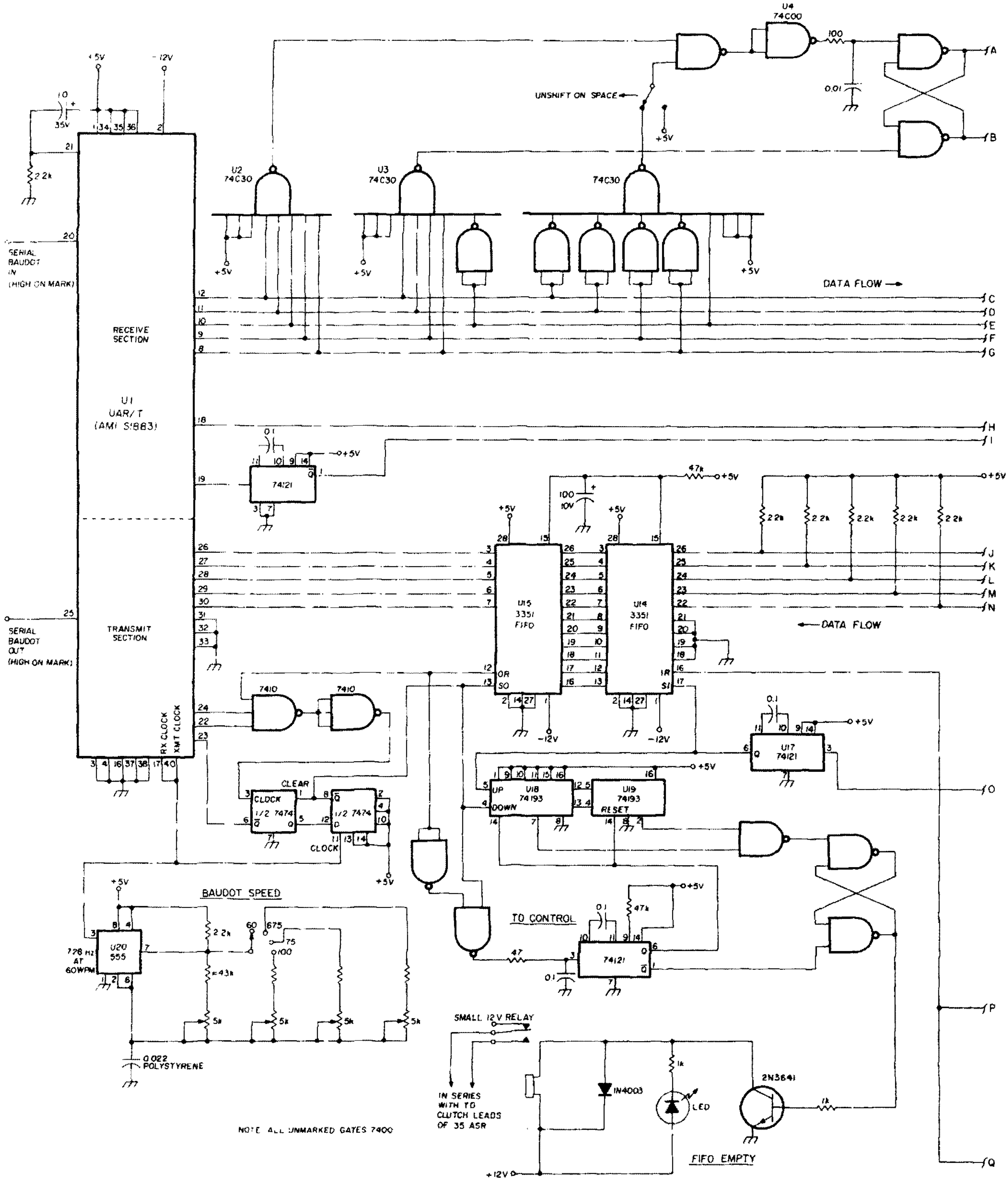



**fig. 1. Schematic of the serial Baudot-to-ASCII-to-Baudot converter for model 33 and 35 ASR machines.**

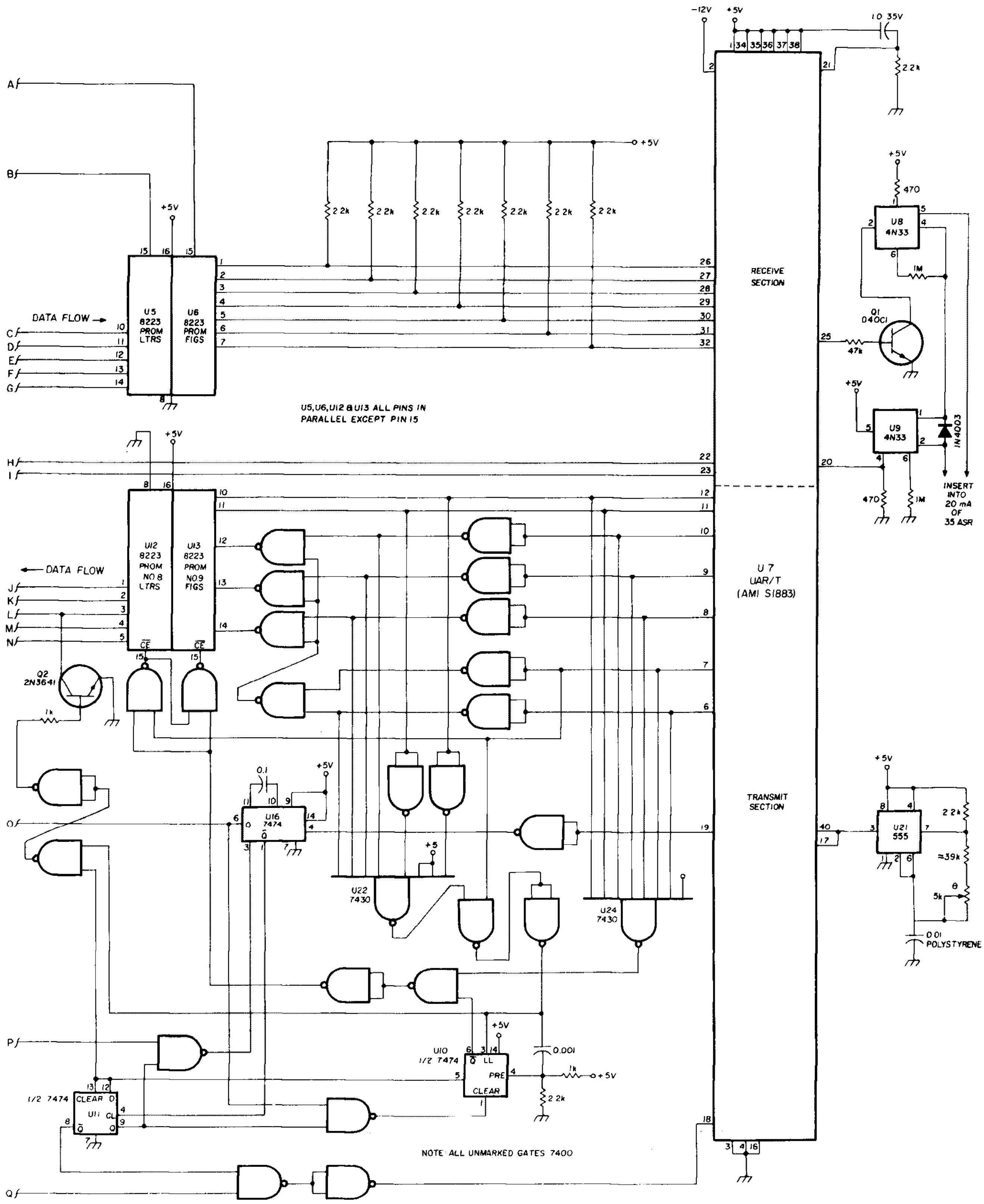

DATA FLOW
U5 8223 PROM LTRS
U6 8223 PROM FIGS
U5,U6,U12 & U13 ALL PINS IN PARALLEL EXCEPT PIN 15
U12 8223 PROM NO 8 LTRS
U13 8223 PROM NO 9 FIGS
Q2 2N3641
U16 7474
U22 7430
U24 7430
U10 1/2 7474
1/2 7474 U11
RECEIVE SECTION
U 7 UAR/T (AMI S1883)
TRANSMIT SECTION
U8 4N33
Q1 D40C1
U9 4N33
1N4003
INSERT INTO 20 mA OF 35 ASR
U21 555
0.01 POLYSTYRENE
NOTE: ALL UNMARKED GATES 7400

table 1. Program for 8223 prom, Baudot-to-ASCII letters (U5).

| Word | A 4 3 2 1 0 | Symbol | B 0 1 2 3 4 5 6 7 |
|---|---|---|---|
| 0 | 0 0 0 0 0 | Blank | 0 0 0 0 0 0 0 0 |
| 1 | 1 0 0 0 0 | T | 0 0 1 0 1 0 1 0 |
| 2 | 0 1 0 0 0 | CR | 1 0 1 1 0 0 0 0 |
| 3 | 1 1 0 0 0 | O | 1 1 1 1 0 0 1 0 |
| 4 | 0 0 1 0 0 | Space | 0 0 0 0 0 1 0 0 |
| 5 | 1 0 1 0 0 | H | 0 0 0 1 0 0 1 0 |
| 6 | 0 1 1 0 0 | N | 0 1 1 1 0 0 1 0 |
| 7 | 1 1 1 0 0 | M | 1 0 1 1 0 0 1 0 |
| 8 | 0 0 0 1 0 | LF | 0 1 0 1 0 0 0 0 |
| 9 | 1 0 0 1 0 | L | 0 0 1 1 0 0 1 0 |
| 10 | 0 1 0 1 0 | R | 0 1 0 0 1 0 1 0 |
| 11 | 1 1 0 1 0 | G | 1 1 1 0 0 0 1 0 |
| 12 | 0 0 1 1 0 | I | 1 0 0 1 0 0 1 0 |
| 13 | 1 0 1 1 0 | P | 0 0 0 0 1 0 1 0 |
| 14 | 0 1 1 1 0 | C | 1 1 0 0 0 0 1 0 |
| 15 | 1 1 1 1 0 | V | 0 1 1 0 1 0 1 0 |
| 16 | 0 0 0 0 1 | E | 1 0 1 0 0 0 1 0 |
| 17 | 1 0 0 0 1 | Z | 0 1 0 1 1 0 1 0 |
| 18 | 0 1 0 0 1 | D | 0 0 1 0 0 0 1 0 |
| 19 | 1 1 0 0 1 | B | 0 1 0 0 0 0 1 0 |
| 20 | 0 0 1 0 1 | S | 1 1 0 0 1 0 1 0 |
| 21 | 1 0 1 0 1 | Y | 1 0 0 1 1 0 1 0 |
| 22 | 0 1 1 0 1 | F | 0 1 1 0 0 0 1 0 |
| 23 | 1 1 1 0 1 | X | 0 0 0 1 1 0 1 0 |
| 24 | 0 0 0 1 1 | A | 1 0 0 0 0 0 1 0 |
| 25 | 1 0 0 1 1 | W | 1 1 1 0 1 0 1 0 |
| 26 | 0 1 0 1 1 | J | 0 1 0 1 0 0 1 0 |
| 27 | 1 1 0 1 1 | FIGS | 0 0 0 1 1 0 0 0 |
| 28 | 0 0 1 1 1 | U | 1 0 1 0 1 0 1 0 |
| 29 | 1 0 1 1 1 | Q | 1 0 0 0 1 0 1 0 |
| 30 | 0 1 1 1 1 | K | 1 1 0 1 0 0 1 0 |
| 31 | 1 1 1 1 1 | LTRS(delete) | 1 1 1 1 1 1 1 0 |

table 2. Program for 8223 prom, Baudot-to-ASCII figures (U6).

| Word | A 4 3 2 1 0 | Symbol | B 0 1 2 3 4 5 6 7 |
|---|---|---|---|
| 0 | 0 0 0 0 0 | Blank | 0 0 0 0 0 0 0 0 |
| 1 | 1 0 0 0 0 | 5 | 1 0 1 0 1 1 0 0 |
| 2 | 0 1 0 0 0 | CR | 1 0 1 1 0 0 0 0 |
| 3 | 1 1 0 0 0 | 9 | 1 0 0 1 1 1 0 0 |
| 4 | 0 0 1 0 0 | Space | 0 0 0 0 0 1 0 0 |
| 5 | 1 0 1 0 0 | # | 1 1 0 0 0 1 0 0 |
| 6 | 0 1 1 0 0 | , | 0 0 1 1 0 1 0 0 |
| 7 | 1 1 1 0 0 | . | 0 1 1 1 0 1 0 0 |
| 8 | 0 0 0 1 0 | LF | 0 1 0 1 0 0 0 0 |
| 9 | 1 0 0 1 0 | ) | 1 0 0 1 0 1 0 0 |
| 10 | 0 1 0 1 0 | 4 | 0 0 1 0 1 1 0 0 |
| 11 | 1 1 0 1 0 | & | 0 1 1 0 0 1 0 0 |
| 12 | 0 0 1 1 0 | 8 | 0 0 0 1 1 1 0 0 |
| 13 | 1 0 1 1 0 | Ø | 0 0 0 0 1 1 0 0 |
| 14 | 0 1 1 1 0 | : | 0 1 0 1 1 1 0 0 |
| 15 | 1 1 1 1 0 | ; | 1 1 0 1 1 1 0 0 |
| 16 | 0 0 0 0 1 | 3 | 1 1 0 0 1 1 0 0 |
| 17 | 1 0 0 0 1 | " | 0 1 0 0 0 1 0 0 |
| 18 | 0 1 0 0 1 | $ | 0 0 1 0 0 1 0 0 |
| 19 | 1 1 0 0 1 | ? | 1 1 1 1 1 1 0 0 |
| 20 | 0 0 1 0 1 | Bell | 1 1 1 0 0 0 0 0 |
| 21 | 1 0 1 0 1 | 6 | 0 1 1 0 1 1 0 0 |
| 22 | 0 1 1 0 1 | ! | 1 0 0 0 0 1 0 0 |
| 23 | 1 1 1 0 1 | / | 1 1 1 1 0 1 0 0 |
| 24 | 0 0 0 1 1 | - | 1 0 1 1 0 1 0 0 |
| 25 | 1 0 0 1 1 | 2 | 0 1 0 0 1 1 0 0 |
| 26 | 0 1 0 1 1 | 1 | 1 1 1 0 0 1 0 0 |
| 27 | 1 1 0 1 1 | FIGS | 0 0 0 1 1 0 0 0 |
| 28 | 0 0 1 1 1 | 7 | 1 1 1 0 1 1 0 0 |
| 29 | 1 0 1 1 1 | ' | 1 0 0 0 1 1 0 0 |
| 30 | 0 1 1 1 1 | ( | 0 0 0 1 0 1 0 0 |
| 31 | 1 1 1 1 1 | LTRS | 1 1 1 1 1 1 1 0 |

If a *LTRS* character must be inserted into the text, transistor Q2, a 2N3641, will be nonconducting so that all bits on the left-hand side of proms U12, U13 will be high, signifying the *LTRS* code. If a *FIGS* character must be inserted into text, Q2 will conduct, making the center bit a zero, which signifies the *FIGS* code.

The ASCII bits are applied to the address lines of 8223 proms U12, U13. Here the ASCII code is translated into the corresponding Baudot code; this data is fed into the two 3351 fifos, U14 and U15. A "data available" pulse at U7-19, delayed by the two 74121 one-shots (U16, U17), appears at the "shift-in" pin of U14 (pin 17). The Baudot characters are thus loaded into the fifo memory.

The fifo memories are necessary because the information from the ASCII machine is fed in at 100 wpm, while the Baudot output from UAR/T U1 is at 60 wpm. When typing, the ASCII speed will exceed 60 wpm only occasionally. However, when the ASCII tape reader runs, the memory will eventually become fully loaded. Because the two fifos can store only 80 characters, the tape reader control circuit, consisting of up-down counter U18, U19, will interrupt current to the tape-reader clutch, holding it until the fifo memory is again empty.

The parallel data at fifo U15 output is applied to UAR/T U1 and appears in serial form at TTL level at U1-25. This signal can be used to key an afsk generator, as described below.

The 555 clocks, U20, U21, will be stable if high-grade components are used. Crystal stability is nice but unncecessary. Precise resistance values have not been given as they may differ from case-to-case, but they can be easily determined with a frequency counter.

Because the equivalent Baudot character for the ASCII space signal is contained in the ASCII-to-Baudot *FIGS* prom, a space signal between words is always preceded and followed by a *LTRS* and *FIGS* signal. This is undesirable because two extra characters will be sent, which are unnecessary. This problem can be resolved by TTL IC U22 which steers U10 (**fig. 1**).

Note that when using the converter with an fsk demoodulator it may be necessary to insert a level changer, inverter, or both between the demodulator and U1-20. The data must enter this point at TTL level. The *Baudot speed* switch (**fig. 1**) allows you to copy signals at speeds other than the amateur speed of 60 wpm.

Pin numbers for the gates in **fig. 1** (type 7400s) have not been given as a new layout will be made for the final version. Except for the proms, the ICs in the Baudot to ASCII section are CMOS devices. These happened to be available, so they were used. The cir-

**table 3. Program for 8223 prom, ASCII-to-Baudot letters (U13).**

| Word | A 4 3 2 1 0 | Symbol | B 0 1 2 3 4 5 6 7 |
|---|---|---|---|
| 0 | 0 0 0 0 0 | Null | 0 0 0 0 0 0 0 0 |
| 1 | 0 0 0 0 1 | A | 1 1 0 0 0 0 0 0 |
| 2 | 0 0 0 1 0 | B | 1 0 0 1 1 0 0 0 |
| 3 | 0 0 0 1 1 | C | 0 1 1 1 0 0 0 0 |
| 4 | 0 0 1 0 0 | D | 1 0 0 1 0 0 0 0 |
| 5 | 0 0 1 0 1 | E | 1 0 0 0 0 0 0 0 |
| 6 | 0 0 1 1 0 | F | 1 0 1 1 0 0 0 0 |
| 7 | 0 0 1 1 1 | G | 0 1 0 1 1 0 0 0 |
| 8 | 0 1 0 0 0 | H | 0 0 1 0 1 0 0 0 |
| 9 | 0 1 0 0 1 | I | 0 1 1 0 0 0 0 0 |
| 10 | 0 1 0 1 0 | J | 1 1 0 1 0 0 0 0 |
| 11 | 0 1 0 1 1 | K | 1 1 1 1 0 0 0 0 |
| 12 | 0 1 1 0 0 | L | 0 1 0 0 1 0 0 0 |
| 13 | 0 1 1 0 0 | M | 0 0 1 1 1 0 0 0 |
| 14 | 0 1 1 1 0 | N | 0 0 1 1 0 0 0 0 |
| 15 | 0 1 1 1 1 | O | 0 0 0 1 1 0 0 0 |
| 16 | 1 0 0 0 0 | P | 0 1 1 0 1 0 0 0 |
| 17 | 1 0 0 0 1 | Q | 1 1 1 0 1 0 0 0 |
| 18 | 1 0 0 1 0 | R | 0 1 0 1 0 0 0 0 |
| 19 | 1 0 0 1 1 | S | 1 0 1 0 0 0 0 0 |
| 20 | 1 0 1 0 0 | T | 0 0 0 0 1 0 0 0 |
| 21 | 1 0 1 0 1 | U | 1 1 1 0 0 0 0 0 |
| 22 | 1 0 1 1 0 | V | 0 1 1 1 1 0 0 0 |
| 23 | 1 0 1 1 1 | W | 1 1 0 0 1 0 0 0 |
| 24 | 1 1 0 0 0 | X | 1 0 1 1 1 0 0 0 |
| 25 | 1 1 0 0 1 | Y | 1 0 1 0 1 0 0 0 |
| 26 | 1 1 0 1 0 | Z | 1 0 0 0 1 0 0 0 |
| 27 | 1 1 0 1 1 | Null | 0 0 0 0 0 0 0 0 |
| 28 | 1 1 1 0 0 | Null | 0 0 0 0 0 0 0 0 |
| 29 | 1 1 1 0 1 | CR | 0 0 0 1 0 0 0 0 |
| 30 | 1 1 1 1 0 | LF | 0 1 0 0 0 0 0 0 |
| 31 | 1 1 1 1 1 | Null | 0 0 0 0 0 0 0 0 |

**table 4. Program for 8223 prom, ASCII-to-Baudot letters (U12).**

| Word | A 4 3 2 1 0 | Symbol | B 0 1 2 3 4 5 6 7 |
|---|---|---|---|
| 0 | 0 0 0 0 0 | Space | 0 0 1 0 0 0 0 0 |
| 1 | 0 0 0 0 1 | ! | 1 0 1 1 0 0 0 0 |
| 2 | 0 0 0 1 0 | " | 1 0 0 0 1 0 0 0 |
| 3 | 0 0 0 1 1 | # | 0 0 1 0 1 0 0 0 |
| 4 | 0 0 1 0 0 | $ | 1 0 0 1 0 0 0 0 |
| 5 | 0 0 1 0 1 | Null | 0 0 0 0 0 0 0 0 |
| 6 | 0 0 1 1 0 | & | 0 1 0 1 1 0 0 0 |
| 7 | 0 0 1 1 1 | ' | 1 1 0 1 0 0 0 0 |
| 8 | 0 1 0 0 0 | ( | 1 1 1 1 0 0 0 0 |
| 9 | 0 1 0 0 1 | ) | 0 1 0 0 1 0 0 0 |
| 10 | 0 1 0 1 0 | Null | 0 0 0 0 0 0 0 0 |
| 11 | 0 1 0 1 1 | Null | 0 0 0 0 0 0 0 0 |
| 12 | 0 1 1 0 0 | , | 0 0 1 1 0 0 0 0 |
| 13 | 0 1 1 0 1 | - | 1 1 0 0 0 0 0 0 |
| 14 | 0 1 1 1 0 | . | 0 0 1 1 1 0 0 0 |
| 15 | 0 1 1 1 1 | / | 1 0 1 1 1 0 0 0 |
| 16 | 1 0 0 0 0 | Ø | 0 1 1 0 1 0 0 0 |
| 17 | 1 0 0 0 1 | 1 | 1 1 1 0 1 0 0 0 |
| 18 | 1 0 0 1 0 | 2 | 1 1 0 0 1 0 0 0 |
| 19 | 1 0 0 1 1 | 3 | 1 0 0 0 0 0 0 0 |
| 20 | 1 0 1 0 0 | 4 | 0 1 0 1 0 0 0 0 |
| 21 | 1 0 1 0 1 | 5 | 0 0 0 0 1 0 0 0 |
| 22 | 1 0 1 1 0 | 6 | 1 0 1 0 1 0 0 0 |
| 23 | 1 0 1 1 1 | 7 | 1 1 1 0 0 1 0 0 |
| 24 | 1 1 0 0 0 | 8 | 0 1 1 0 0 0 0 0 |
| 25 | 1 1 0 0 1 | 9 | 0 0 0 1 1 0 0 0 |
| 26 | 1 1 0 1 0 | : | 0 1 1 1 0 0 0 0 |
| 27 | 1 1 0 1 1 | ; | 0 1 1 1 1 0 0 0 |
| 28 | 1 1 1 0 0 | Null | 0 0 0 0 0 0 0 0 |
| 29 | 1 1 1 0 1 | Null | 0 0 0 0 0 0 0 % |
| 30 | 1 1 1 1 0 | Null | 0 0 0 0 0 0 0 0 |
| 31 | 1 1 1 1 1 | LTRS | 1 0 0 1 1 0 0 0 |

cuit should work just as well with TTL devices. **Fig. 2** shows socket connections for the devices.

The circuit in **fig. 3** makes programming a cinch as the burn-out time is automatically determined by the 74121 one-shot, U1. The time is fixed at 150 milliseconds. Programming must be done carefully, while you're wide awake, or mistakes are bound to occur! The +15 and +5-volt leads must be connected to a regulated power supply that provides at least 1 ampere. Proceed as follows:

1. With the power supply shut off, insert the 8223 to be programmed into its socket.

2. Set S2B to *BURN*.

3. Set address switches S3-S7 and output switch S8 according to the program pattern for the first bit.

4. Switch on the power supply and depress S1.

5. Set S3-S7 and S8 to the next bit and depress S1. Continue this procedure, bit-by-bit, until the entire pattern is programmed into the chip. You can test the programming by setting S2 to *TEST*. Go through the entire pattern again, using switches S3-S7 and S8. The LED will illuminate for a 1 and remain dark for a zero. If the test yields the desired pattern your 8223 is ready for use.

Four 8223s are necessary for the Baudot-to-ASCII conversion and vice versa. For more information on the makeup of the programming pattern, see reference 3. An article describing a memory for automatic CW identificaton using an 8223 prom can be found in reference 4.

In **tables 1-4** you'll find one program for each of the four proms used in the code converter. The *A* column determines the prom address line switch positions, while the *B* column determines the switch positions of the prom outputs.

## example

To program the letter *Y* into the prom that translates Baudot to ASCII letters (**table 1**), proceed as follows:

1. Set switches S2A-B to *TEST* and set the prom output line-selector switch, S8, to *B0*.

2. Look up the letter *Y* on the program (**table 1**).

3. Set the address line switches, S3-S7 (**fig. 3**) according to the information in the table: *A4* = high; *A3* = low; *A2* = high; *A1* = low; and *A0* = high — i.e., 10101. (Low is ground and high is +5 volts.)

4. Set switches 2A-B to *BURN*.

5. Set the prom output line switch to *B0* (a 1 in this case), then depress switches S1A-B.

6. Advance the output line selector switch to *B3*, picking up another 1.

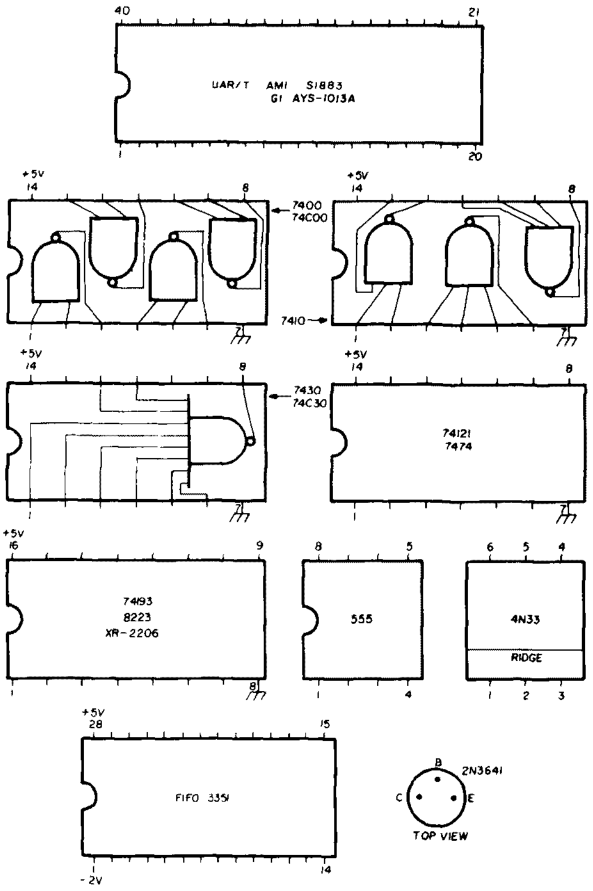



fig. 2. Socket connections (top view) for the devices used in the converter.

7. Press switches S1A-B again.

8. Repeat the procedure at *B4* and *B6*, picking up a 1 in each case.

9. Set switches S2A-B to *TEST*. With S8 in *B0, B3, B4,* and *B6* positions, the LED should light, indicating a 1; in positions *B1, B2, B5,* and *B7* it should remain dark, indicating a low, or zero. Should one of the prom fuses refuse to open at the first attempt at testing, try burning it repeatedly.

Start the programming from the top of the table and work down,. When you have one prom fully programmed, go through the check procedure with switches S2A-B in the *TEST* position to verify that the fuses have opened according to the program in use. If the prom checks out okay, mark it for identification.

Type 82S23 proms are presently offered on the surplus market at a reasonable price. These can also be used, but to program these devices the 390-ohm ½-watt resistor at the top contact of S2A (**fig. 3**) must be changed to 4.7 ohms ½-watt, and the +15-volt supply must be increased to 16 volts.

In reference 5, I pointed out the importance of phase-continuous frequency shift to prevent transients, which can cause interference on adjacent channels. I described a rather commplex circuit of an afsk generator that featured phase continuity of the sine-wave output when keyed.

In the meantime integrated circuits have appeared on the market especially designed for waveform generation. One of these, the EXAR XR-2206 CP, is particularly interesting for RTTY. Pin 9 of this IC can be directly connected to pin 25 of U1, **fig. 1**, of the Baudot-to-ASCII/ASCII-to-Baudot converter. The output at pin 2 of the XR-2206 can then be used to modulate an ssb transmitter or transceiver with a phase-continuous afsk signal.

The afsk circuit (**fig. 4**) is extremely simple. The sine-wave frequency is determined by the value of C1 and the total resistance connected to either pin 7 or 8 of the XR-2206. When pin 9 is high pin 7 is active; when pin 9 is low pin 8 is active.

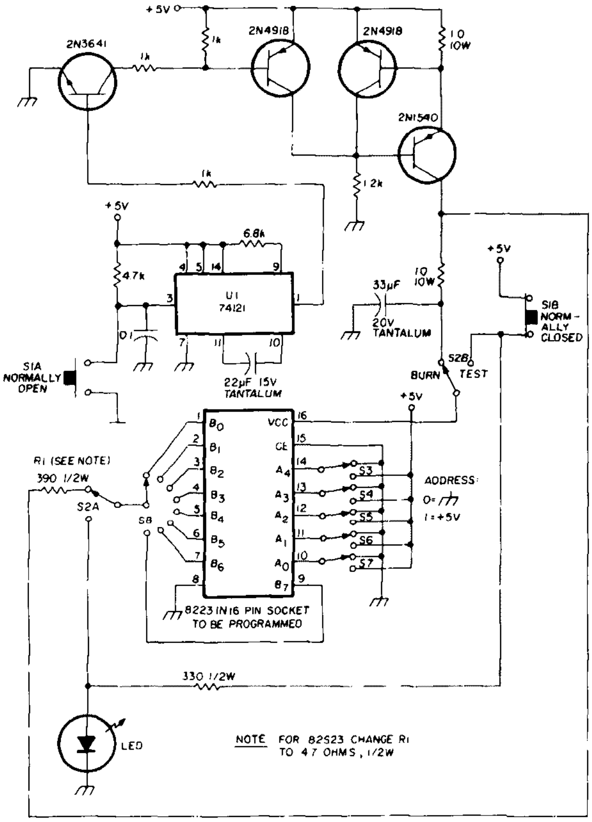



fig. 3. Circuit for programming the 8223 proms.

For excellent frequency stability, C1 should be a polystyrene (Phillips 295 AA/C 8K2) or a Mylar capacitor. The frequency-determining resistors should be carbon film; or better yet, metal film types.

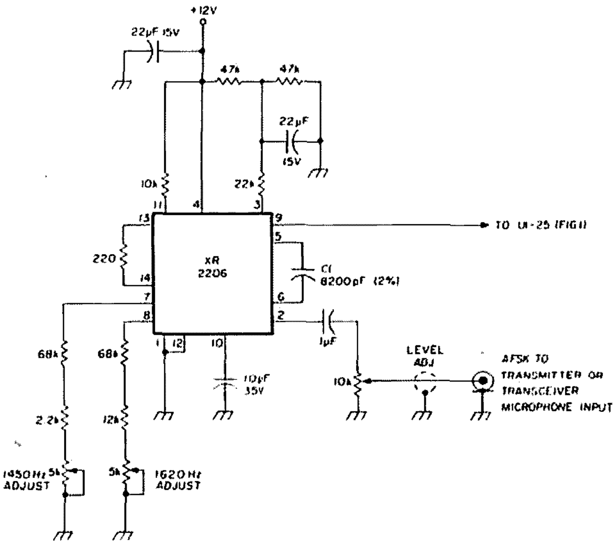



fig. 4. AFSK generator. Phase-continuous frequency shift is featured to prevent out-of-band transients. Sine-wave frequency is determined by C1 and the total resistance connected to either pin 7 or 8 of the XR2206.

The afsk-generator audio frequencies of 1620 and 1450 Hz were chosen to put the second harmonic outside the passband of modern ssb equipment and to eliminate the need for special carrier-frequency crystals in such equipment.

The overall converter system can be tested by feeding data from a Baudot keyboard or tape reader to U1-20 (**fig. 1**). As the output and input of U7 are a closed loop through the two 4N33 optoisolators, U8 and U9, the data is fed back to U1, and its output at pin 25 can be used to operate the printer magnets of the Baudot machine through a suitable driver. In this case, the Baudot data is converted to ASCII, then back to Baudot. The only character not translated from ASCII to Baudot is the bell signal. With additional gates this could be accomplished, but I felt that the additional complexity was unjustified.


## acknowledgement

I'd like to thank my friend, Paul Hudson, VE3CWA, for suggestions in preparing this article.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**ham radio**

# admittance, impedance and circuit analysis

**A circuit with impedances** seems very complicated to those lacking experience in circuit analysis. Throw in names like *admittance*, *conductance*, or *susceptance*, and circuit analysis seems to take on a certain mystique beyond the average amateur. Many texts present such a bewildering array of symbols and mathematical operators that analysis seems hopeless unless you have a computer. Wrong – all you need is paper, pencil, and a pocket calculator.

Impedance, admittance, and complex numbers are really simple expressions with a few extra rules. Once you know them and how to set up an analysis "model," the rest is just careful number manipulation. Presented here are the basics, how to handle them, and simple circuit modeling for the majority of transmitter and receiver circuits.

If the names and expressions are unfamiliar, just read and study slowly. They will become familiar with a little practice.

## the rectangular form

The two forms of expressing impedance or admittance are *rectangular* and *polar*. **Fig. 1** shows the rectangular form with the expressions and circuit symbols in columns for impedance, $Z$, and admittance, $Y$.

Any L C R (inductance, capacitance, and resistance) combination can be shown as either an impedance or an admittance because each is the inverse of the other. Common use has a series combination given as impedance, a parallel combination as admittance. Opposite expressions are shown later.

If you are unfamiliar with complex number notation, pay attention to the $j$ symbol. The $j$, also called the *j-operator*, denotes everything to the right of it as *imaginary*. This is in the mathematical sense since $j = \sqrt{-1}$, an imaginary number. A complex number has a *real* part (the resistance, $R$, or conductance, $G$) and an *imaginary* part (the reactance, $X$, or susceptance, $B$). Ordinary math rules apply only to the real part or the imaginary part. Rules for handling both at the same time are given later.

Values of impedance, resistance, and reactance are given in familiar ohms values. Values of admittance, conductance, and susceptance are given in *mhos*. Some time ago the inverse spelling was applied because admittance is the inverse of impedance. If $R = 10\ ohms$, then $G = -0.1\ mho$.

## reactance and susceptance

Both are frequency sensitive; that is, their ohm and mho values vary with frequency. This is shown by

$$X_L = \omega L \qquad B_L = -1/\omega L$$
$$X_C = -1/\omega C \qquad B_C = \omega C$$

with L in henries, C in farads, and

$$\omega = 2\pi f$$

where f is in hertz (radian frequency). The subscript refers to inductance or capacitance. Total reactance or susceptance is expressed without a subscript as

$$X = \omega L - (1/\omega C)$$
$$B = \omega C - (1/\omega L)$$

Note especially that signs are shown; this *must* be followed in reactance and susceptance calculation. Remember the resonance condition where inductive and capacitive reactance (or susceptance) cancels.

Imaginary parts can be stated in opposite terms:

$$X = -1/(B_C + B_L) = -1/B$$
$$B = -1/(X_C + X_L) = -1/X$$

Note the signs. Reactance and susceptance are related by *negative* inversions while resistance and conductance are related by *positive* inversions. Real and imaginary parts are handled separately andd the part relationships are different but admittance is still the inverse of impedance and vice-versa. $Y$ and $Z$ are complex numbers of different rules while $R$, $X$, $G$ and $B$ are simple numbers with ordinary rules.*

Most modern hand-held calculators have a pi-constant key and at least one memory register. Storing the radian frequency ($2\pi f$) in memory allows rapid calculation of $X$ or $B$ values. For rf circuits, entering scaled values of megahertz, microhenries,

*A good algebra text will show the proof and following rules for those readers desiring more information.

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or microfarads saves using the exponent key or inputing lots of zeros.

With complex numbers $(a+jb)$ and $(c+jd)$, the rules are:

$$(a+jb)+(c+jd)=(a+c)+j(b+d)$$
$$(a+jb)-(c+jd)=(a-c)+j(b-d)$$

Notice the expression statements on each side. On the left there are two complex numbers indicated by the parenthesis. The right side shows that each part of the complex answer has at least two terms. The $j$-operator designates the entire $b$, $d$ term group as imaginary.

The addition rule is useful for expressing a combination LCR circuit as one impedance if all components are in series. Suppose the inductor has winding resistance. The total resistance is the sum of $R$ and winding resistance. The inductive reactance adds to capacitive reactance and each could be in the two impedances or combined in one of them. Any number of combinations, including parallels, can be added to give one real part and one imaginary part. Remember to keep the parts separate, observe signs, use admittance for parallel combinations, and impedance for series combinations.

**table 1. Keyboard steps, HP-35 calculator impedance/admittance conversion.**

| | Keyboard | Stack Registers x | y | z | t | Display, (remarks) |
|---|---|---|---|---|---|---|
| | Input R | R | | | | |
| 1 | STO | R | | | | (hold R in memory) |
| | Input X | X | | | | |
| 2 | ENTER | X | X | | | |
| 3 | ENTER | X | X | X | | (fill stack with X) |
| 4 | × | $X^2$ | X | | | |
| 5 | RCL | R | $X^2$ | X | | |
| 6 | RCL | R | R | $X^2$ | X | |
| 7 | × | $R^2$ | $X^2$ | X | | |
| 8 | + | $Mag^2$ | X | | | |
| 9 | $\sqrt{x}$ | Mag | X | | | Z Magnitude |
| 10 | 1/x | 1/M | X | | | Y Magnitude |
| 11 | xy | X | 1/M | | | |
| 12 | RCL | R | X | 1/M | | |
| 13 | ÷ | X/R | 1/M | | | |
| 14 | arc | | | | | |
| 15 | tan | pha | 1/M | | | Z phase angle |
| 16 | CHS | − pha | 1/M | | | Y phase angle |
| 17 | STO | | | | | (hold angle in memory) |
| 18 | R↓ | 1/M | | | | |
| 19 | ENTER | 1/M | 1/M | | | |
| 20 | ENTER | 1/M | 1/M | 1/M | | (fill stack with Y magnitude) |
| 21 | RCL | − pha | 1/M | 1/M | 1/M | |
| 22 | cos | Cos | 1/M | 1/M | 1/M | |
| 23 | × | G | 1/M | 1/M | | conductance |
| 24 | x↔y | 1/M | G | 1/M | | |
| 25 | RCL | − pha | 1/M | G | 1/M | |
| 26 | sin | Sin | 1/M | G | 1/M | |
| 27 | × | B | G | 1/M | | susceptance |
| 28 | 1/x | 1/B | G | | | |
| 29 | CHS | Xp | G | | | parallel reactance |
| 30 | x↔y | G | Xp | | | |
| 31 | 1/x | Rp | Xp | | | parallel resistance |

The sign in front of the $j$-operator can take the sign of the imaginary part. This is strictly a sign and *does not* mean the *parts of one* complex number add or subtract. The subtraction rule will be useful in impedance matching to be discussed later.

## why admittance?

There are as many parallel combinations of com-

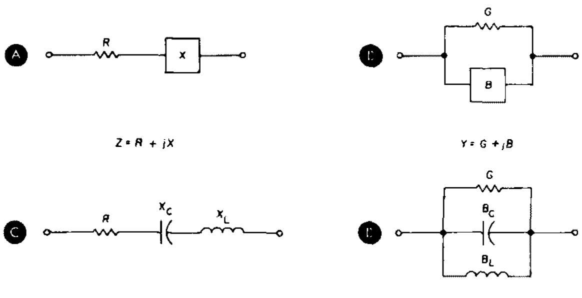



**fig. 1. Rectangular form of impedance and admittance.**

ponents as there are those in series. Any circuit to be analyzed should be separated into *branches* of component combinations in series or parallel. Each branch is then expressed as an admittance or impedance having only two connections or *nodes*. The easiest form of expressing a parallel combination is admittance since all real parts and all imaginary parts add.

In fact, you have probably been using admittance without realizing it. The familiar parallel resistance formula

$$R_{total} = \frac{R_1 R_2}{R_1 + R_2}$$

was derived from the basic expression

$$R_{total} = \frac{1}{(1/R_1) + (1/R_2) + \cdots (1/R_n)}$$

Invert the basic expression and you will find the sum of conductances $(1/R)$ equals total conductance.

## multiplying and dividing complex numbers

These will be used in solving the circuit or converting admittance to impedance and vice-versa. The rules are:

$$(a+jb) \times (c+jd) = (ac-bd)+j(ad+bc)$$

$$\frac{(a+jb)}{(c+jd)} = \left[\frac{(ac+bd)}{(c^2+d^2)}\right] + j\left[\frac{(bc-ad)}{(c^2+d^2)}\right]$$

$$(inversion)\, 1/(a+jb) = [a/(a^2+b^2)] - j[b/(a^2+b^2)]$$

These rules are hard to work with, even with a calculator. There is an easier way by using the *polar form equivalents* shown in **fig. 2**. The angle symbol is as important as the $j$-operator: All terms to the

right of the angle symbol are, appropriately, *angles;* terms to the left are *magnitudes*.

$$(A \angle \phi) \times (B \angle \theta) = (A \times B) \angle (\phi + \theta)$$

$$\frac{(A \angle \phi)}{(B \angle \theta)} = \frac{A}{B} \angle (\phi - \theta)$$

$$(inversion) \quad \frac{1}{A \angle \phi} = \frac{1}{A} \angle - \phi$$

Magnitudes operate algebraically just like the polar expression, but angles add or subtract. Again, the

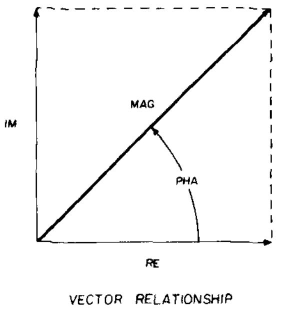



**fig. 2. Comparison of the rectangular and polar form.**

parts are different from the whole. A calculator with trig functions is preferred; one with built-in polar/rectangular conversion is even better.

## polar form

$Y$ and $Z$ can be expressed equally well this way. This has the advantage of showing magnitude and phase angle as shown by the little vector diagram of **fig. 2**. Voltage and current in a complex circuit will also be complex quantities. The ac or rf voltage you measure in a circuit is the magnitude. Phase angle is also measurable although the instruments cost more. Some rf bridges read directly in polar form.

Both forms are needed in circuit analysis. Rectangular form can be used in calculating circuit branch values while the polar form is used for voltage response and as an intermediate step in conversion. Input impedance or admittance can be expressed in either form. The angle is a signed number but the magnitude is always a *positive* number.

## form conversion by calculator

All of the expressions and forms necessary for circuit analysis have now been presented. Before going into the analysis setup it is useful to observe conversion of impedance to admittance with the aid of rectangular/polar form changes. The program steps in **table 1** apply to the older HP-35 model or any other *RPN/Stack* machine that does not have polar/rectangular conversion functions.

This form conversion is basic to the ladder network solution to be presented later. Calculators with built-in functions can use part of this step sequence, either manually or automatically with a programmable version.

**Steps 4** through **15** change rectangular impedance to polar impedance. **Step 10** finds the polar admittance magnitude; its location in the sequence is arbitrary and could be placed after **step 18. Step 16** finds the *polar admittance angle*.

**Steps 17** to **20** set up the registers for polar to rectangular conversion — the remaining steps perform it. The magnitude inversion and angle sign change has already done conversion of impedance to admittance. Polar/rectangular identities are the same for impedance and admittance.

**Steps 28** to **31** are used only to show how parallel resistance and reactance are derived from the main program. There are simpler ways to derive the parallel equivalents from series components. You are invited to try out the simpler way from the formulas given. If you succeed, you have already begun programming.

**Steps 1** to **27** will convert admittance to impedance without changing any steps. To prove this to yourself, just change the terms and expressions in the register and remarks columns.

As an example, $Z = 4 + j3$ *ohms* converts to $Y = 0.16 - j0.12$ *mhos*. The parallel reactance equivalent is **8.33** ohms, resistance equivalent is **6.25** ohms.

## the analysis model

The term *model* simply means the branches connected at nodes. A reduction to this structure is done to simplify overall calculations calculation time and effort. The basic pi structure in **fig. 3** can be used with almost every circuit and is expandable to a ladder structure.

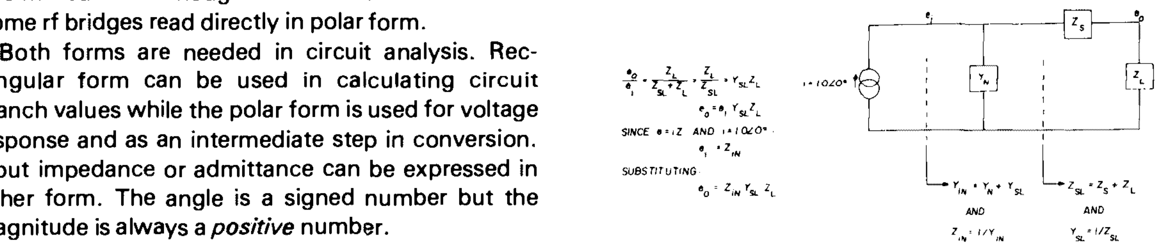



**fig. 3. Basic model structure of the familiar pi network.**

The model is allowed only one signal source, a current of 1.0 ampere with no phase angle. Any source admittance must be included as part of $Y_n$. A unity current source allows each output voltage, $e_o$, to be relative and the solution to be dependent solely on impedances and admittances in the model. This is the simplest method and applies well to calculator solution.

It may seem that a relative $e_o$ solution is unrealistic

but this is only partly true. Networks such as filters are almost always described in relative terms. As an example, an i-f amplifier can be modeled as several individual blocks with individual block model response summed for overall response. The actual $e_o$ can be found by multiplying relative $e_o$ by the actual input current.

Most of us think of an amplifier stage as output vs input. For the model the transistor or tube must be split into the input impedance and the output admittance with a current source that is dependent on input signal. Most data sheets on transistors and tubes are oriented to this form.

Most of us also think of an amplifier output in terms of voltage. Since voltage is easiest to measure, it is easy to neglect current. All transistor collectors and pentode tube plates are really current sources with an output admittance ($h_{oe}$ for common emitter, $1/r_p$ for pentodes). Measured output voltage is the product of the current source and the *total* load impedance, including output admittance of the device. The scarcity of ac current measuring instruments has misled a lot of us in understanding actual circuit operation.

The basic pi model can be thought of as the coupling circuit between two amplifiers. The output admittance of the first becomes part of $Y_n$, and the input impedance of the second becomes part of $Z_L$. The current flowing into the second state input can be found relative to unity (the model input current) because the output voltage is known.

You can use Ohm's law with complex quantities. The difference from dc is that current and voltage have *both* magnitude and phase angle. The rectangular form also applies, but the polar form is more convenient since an ac voltmeter measures magnitude. An oscilloscope will display voltage phase angle but that is dependent on the scope sweep triggering point; a dual-trace scope shows *relative* voltage phase.

Note the progression of $Z$ and $Y$ looking towards the load in **fig. 3**. This is important to the model extension following.

## extending the model for ladder networks

A *ladder* configuration is a shunt-series-shunt sequence — a five-branch model is shown in **fig. 4**. Again, the current source is unity and $e_o$ is relative and solved directly from impedances and admittances. This model is well suited to filter analysis.

Output voltage solution is self-explanatory but note the progression of $Y_{12}$ to $Z_{45}$.* $Y_{12}$ is the inverted sum of $Z_d$ and $Z_L$; $Z_{23}$ is the inverted sum of $Y_c$ and $Y_{12}$, and so on. Each $Y$ or $Z$ is the total, looking towards the load, of the model $Y$ and $Z$ at each node. $Z_{45}$ is the *total* input impedance; the network input admittance is found by subtracting source admittance from $Y_{45}(1/Z_{45})$.

So far, circuit analysis seems complicated and a lot of work. Here is where the calculator comes in handy and it is useful to recall the form-conversion calculator steps in **table 1**. Let's assume that all shunt branches have been precalculated for admittance and all series branches in impedance.

Enter $Y_L$ and begin the conversion. Pause when you reach the magnitude and angle of $Z_L$ and write down the polar values. By **step 27** the rectangular

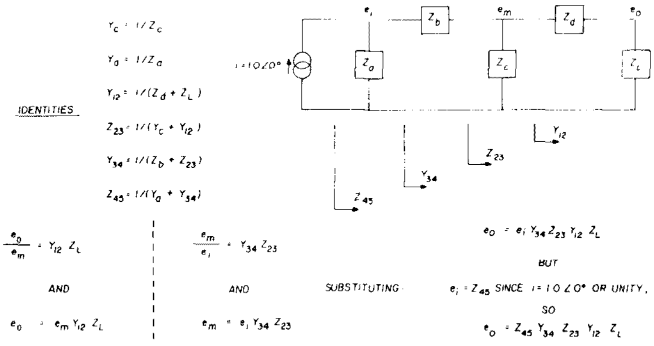



**fig. 4. Five-branch ladder network, showing derivation of circuit quantities. A hand-held calculator is very useful in solving problems of this type.**

form of $Z_L$ is reached and the real and imaginary parts of $Z_d$ are added. Go back to the beginning and start the conversion for $Y_{12}$; the calculator's stack already has the entry so it is just a matter of storing the real part in memory and the imaginary part in the stack. Pause at the polar form of $Y_{12}$ and write down the values. At the end of the second run add the precalculated $Y_c$ parts to parts of $Y_{12}$ and return to the beginning. Continue the repetition, writing down each numbered-subscript polar value, until the input branch has been reached.

Solution of $e_o$ is made by multiplying all tabulated magnitudes and adding all tabulated angles. This is the same as the last $e_o$ expression in **fig. 4** and polar multiplication rules are used. Network input impedance can be found by subtracting source admittance using rectangular form rules and converting.

Six-digit accuracy of precalculation and tabulation is more than sufficient for most purposes; angle tabulation can be rounded to three fractions using degrees, four with radians. If a mistake is made in stepping through the sequence, just begin in mid-

*Amateurs who are not familiar with networks are often confused by the numeric subscripts which are used. The designator $Y_{12}$, for example, refers to the admittance looking into sections 1 and 2 of the network; likewise, $Z_{45}$ refers to the impedance looking into sections 4 and 5 of the network (see fig. 4). In filters, capacitors and inductors are often designated in the same manner — in terms of their relative positions within the filter network.

sequence by clearing registers and entering the last tabulated polar value; just make certain that magnitude and angle are positioned correctly.

The repetitive sequence or iteration is well suited to a programmable calculator. If it has a larger memory, the partial products of $e_o$ can be accumulated automatically.

## modifying a model with branches that jump adjacent nodes

**Fig. 5** shows this condition and defeats the ladder configuration. **Fig . 5** also shows how a model subsection can be transformed by the *delta-to-tee* method, yielding a result that fits the ladder. Transformation must be done with precalculated values.

A delta sub-section that is pure resistance requires transformation only once. Any other condition requires a transform at every solution frequency. The resulting tee values become the new precalculated values for solution iteration.

Similar transformations can be made for tee-to-delta and lattice networks. Most engineering handbooks contain the transform equations, usually given as impedances, but the equations work just as well with admittances. It is worthwhile to study the circuit to be modified and arrange the branches for the least amount of transformation.

Pi networks are good examples and published design data can be used as a starting point. Unfortunately, most designs assume only a resistive load while an actual load such as an antenna will vary considerably. The object of impedance matching is to make the real part of the matched load equal to the source, and reduce the imaginary part to zero.

Let's take the case of matching an antenna over a few frequencies with a pi network. The basic design data is available and the components can be modeled as in **fig. 3**. The approximate antenna impedance data is known either by measurement or from handbook data. Will the network components do the job?

A way to find out is to solve only for $Z_{in}$, *omitting* source admittance and susceptance of the first-branch network componentt (usually a variable capacitor). The rectangular form of $Z_{in}$ is a bit better for solution.

Tabulate the results at each frequency and antenna impedance expected and examine the imaginary part sign. If the sign does *not* change, the input component can be reactive and the value can be calculated by the *opposite-sign* reactance. The imaginary part must go to zero when the component is included in the model. Value range variation can also be seen; make certain that this can be realized in practice.

An imaginary-part sign change means that another component value must change. The most common

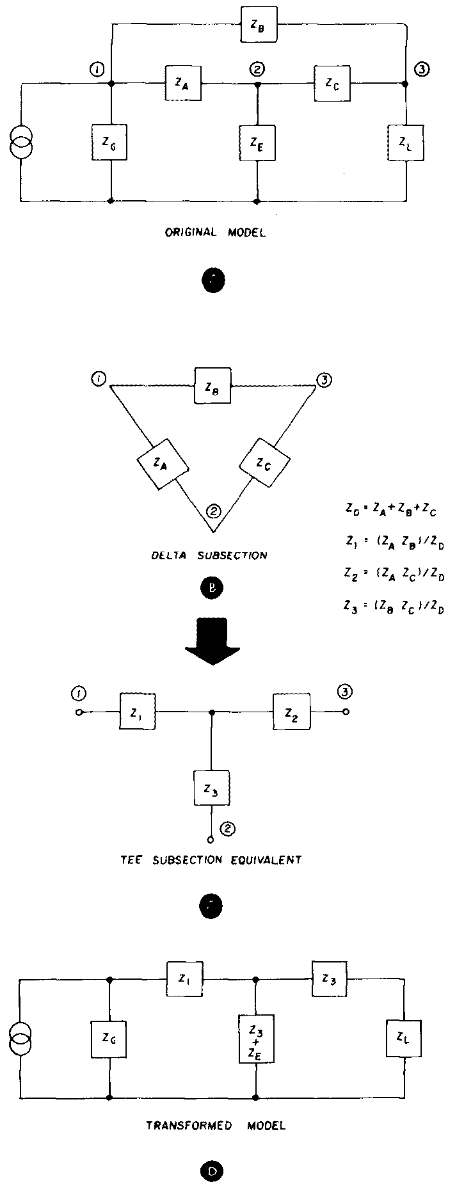



**fig. 5. Model transformation. Here the delta subsection (consisting of $Z_A$, $Z_B$, and $Z_C$) is transformed into an equivalent tee section. The tee section is then added to $Z_G$, $Z_L$, and $Z_O$ to yield the transformed model.**

pi network uses a variable capacitor at each end with a fixed series inductor. If this is the circuit, try varying the other capacitor, solving again and inspecting the signs.

The real part of $Z_{in}$ should be reasonably close to the required resistive value for best power transfer. The difference depends on the type of source. A 20 to 30 per cent variation is probably good enough in most cases. If this part is not close, another component value must be changed. After a few changes you will be able to tell which way to change and you can zero in on the correct set of values. The best effi-

ciency occurs with source impedance equal to matched load impedance.

Other networks can be designed from the efficiency rule and reducing reactance to zero. Some careful study and algebra will result in a matching formula. Impedance or admittance — as a quantity — obeys algebraic rules. The only difference is that real and imaginary parts must be handled by the rules given.

## circuits with more than one load

Suppose you have a circuit model like **fig. 4** and there are two $Z_d$ and two $Z_L$ branches in separate paths. Solve $Y_{12}$ for the second path and record it. Solve the first path as before except the second path's $Y_{12}$ is also added to $Y_c$ plus $Y_{12}$ of the first path. $Z_{23}$ is the total impedance of both paths.

The second path $e_o$ is found from $Z_{45}$, $Y_{34}$, $Z_{23}$ recorded in solving the first path and $Y_{12}$ and $Z_L$ of the second path. This can be extended to longer paths with different lengths provided that the common-node impedance value is the total impedance of all paths.

## bilaterality

All illustrations have shown the source at left, load at right. This conforms to conventional left-to-right flow but doesn't mean the schematic must be interpreted this way. Many schematics are drawn differently, so take care in forming the model — properly locate the source and load.

## Q equivalents

Every coil and capacitor is lossy. To properly analyze a filter this loss must be modeled as resistance or conductance in the proper branches. General values are $X/Q$ for impedance and a series resistance, $B/Q$ for admittance and a parallel conductance. Add losses in LC branches.

Some simplification is possible. $Q$ is fairly constant over an octave of frequency. A fixed $R$ or $G$ value can be used for resonant circuits and filters. The fixed value would be obtained at the center frequency or cutoff frequency for highpass and lowpass filters.

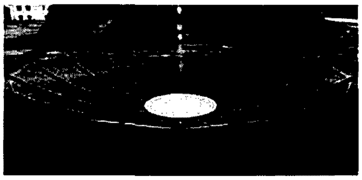
## other analysis methods

Most circuit theory texts have them. The inexperienced should be cautious since the math is high level, usually involved with matrices or transfer functions. A matrix is best solved on a computer. The other methods are more versatile, but not more accurate. The ladder form given here will fit a manual or programmable pocket calculator better. It will not solve all circuits, but most of them. Programs for the HP-25 will be given in a future article.

**ham radio**

# NEW products

## parabolic dish reflector kit



Moonbouncers, radio astronomers, TV DXers, and others who have needed a parabolic dish antenna at a reasonable price will like the *Paraframe* — a rigid, accurate, parabolic antenna framework kit which one person can easily assemble with ordinary hand tools. The *Paraframe* design permits the reflector to be fine-focused for peak performance, yet the weight is low, while strength is high.

Prestressed *Paraframes* are made of wood, and the ribs are furnished with two coats of premium-quality latex exterior sealer-primer, followed by two coats of premium quality latex exterior paint in a neutral gray. Other colors are available upon request.

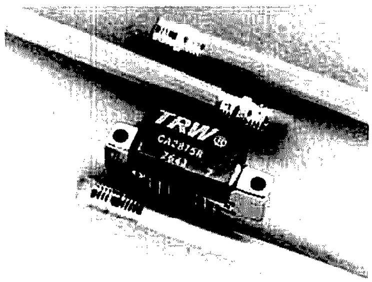
The *Paraframe's* price is such that you can build a parabolic reflector for less than one-third the cost of a similar antenna of equivalent aperture. For example, the SD12 is a 12-foot (3.66-meter) design with a focal length/diameter ($f/d$) ratio of 0.5. Sixteen prestressed ribs provide sufficient rigidity for useful work at 2300 MHz in a 40-mph (65-kmh) wind. The ribs will accept 7/16 inch (11mm) staples for attaching the aluminum window-screen reflector material. Model SD16 has a sixteen foot (4.88 meter) diameter and 20 ribs; model SD20 has a twenty-foot (6.1 meter) diameter and 24 ribs.

A complete selection of expanded and perforated aluminum covering material, together with low-loss coaxial cable suitable for uhf purposes, is available from the *Paraframe* kit supplier.

The model SD12 is priced at $690 (shipping weight is 140 pounds or 64 kg); model SD16 is $1120 (shipping weight is 230 pounds or 104 kg); and model SD20 is priced at $1680 (shipping weight 400 pounds or 181 kg). Shipping cost will be approximately $25 per cwt (45kg), maximum, anywhere in the continental United States. Terms are 50% of the purchase price with the order, the balance COD. Illinois residents are asked to include 5% sales tax.

For a complete description and illustrated brochure, write to James K. Vines, 611 Farmview Road, Park Forest South, Illinois 60466, or call (312) 534-0889 after 7 PM CDT.

## rf hybrid amplifiers




TRW RF Semiconductors has introduced an rf-hybrid gain block which will meet or exceed the most demanding requirements of i-f amplification in advanced microwave radio relay system applications.

The amplifier, designated CA2875/2875R, has a noise figure of typically 4 dB and a third-order intercept of +42 dBm. Requiring a 15 to 24 volt power supply, the CA2875 is suitable for positive power supply polarity while the CA2875R will accommodate a negative supply polarity.

Other parameters of the hybrid amplifier include a return loss of greater than 30 dB at both the input and output ports, phase linearity from 30 to 110 MHz and wide dynamic range. These i-f gain/blocks have a center frequency of 70 MHz as well as a nominal gain of 17.5 dB and an operating temperature range of −40°C to +100°C.

In quantities of 100 pieces, the CA2875/2875R is priced at $31.50. For more information contact Warren Gould at (213) 679-4561 or TRW RF Semiconductors, 14520 Aviation Boulevard, Lawndale, California.

## equipment directory

Have you ever searched through a pile of magazines, a loose-leaf collection of product releases, or a collection of dog-eared catalogs looking for a particular antenna, the specs on a new transceiver, or the nearest distributor-dealer for a certain brand of amateur equipment, only to give up in frustration?

Well, you don't have to repeat that futile exercise this year. The new *1977 Amateur Radio Equipment Directory*, published by Kengore Corporation, got it all together just for you. Here is a comprehensive catalog of amateur equipment, complete with the names and addresses of manufacturers and distributors, together with product photographs, specifications, and prices, conveniently and

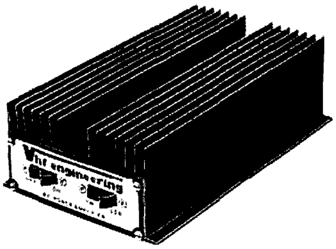


More Details? CHECK—OFF Page 126

attractively bound between soft covers for your reference library.

Not every last item made by every manufacturer is listed, nor do the prices reflect recent price increases, but the catalog lists telephone numbers where you can get the latest, correct information. In spite of these minor (and expected) shortcomings, you'll have to look a long time before finding anything nearly as useful or informative. For your copy, send $2.95 to Kengore Corporation, 9 James Avenue, Kendall Park, New Jersey 08824.

## 70-watt two-meter amplifier



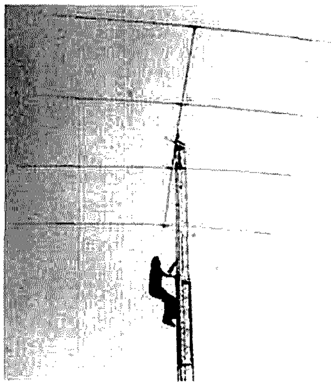
A new 70-watt, four-mode, two-meter amplifier has been introduced by VHF Engineering. This new amplifier, the Blue Line BLC 10/70, is designed to be used with the popular 10-watt fm transceivers and multi-mode transceivers in the 5-15 watt class; it will deliver 70 watts output in both the class C or linear mode. An additional model, the Blue Line BLC 2/70, offers the same features as the BLC 10/70 but will operate with transceivers or transmitters in the 1 or 2 watt class.

The VHF Engineering Blue Line series of amplifiers have been designed for reliability and long life and feature unique broadband, strip-line designs which require no tuning or adjustment during their lifetime. Automatic sensing and relay switching are provided to automatically switch the amplifier into the circuit when drive is applied in the class C (fm) or linear (ssb) modes. The amplifiers offer high efficiency and introduce a receive insertion loss of less than 1 dB. They are designed for

12-14 Vdc operation in base station or mobile service.

The BLC 10/70 sells for $139.95 and the BLC 2/70 sells for $159.95. These new four-mode amplifiers are available from dealers nationwide or from VHF Engineering, 320 Water Street, Binghamton, New York 13902, as wired and tested units.

## new antennas from cubic corporation



A series of new amateur radio antennas, including four beam, two mobile, and one trap vertical model, are available now from Swan Electronics, a subsidiary of Cubic Corporation. The fixed antennas include the TB-4HA, a triband beam for $259.95, featuring four working elements on 10, 15, and 20 meters; the 24-foot boom permits optimum spacing for maximum forward gain and front-to-back ratio.

Also available are the TB-3HA, a triband beam for $199.95 which features three working elements on 10, 15, and 20 meters with a 16-foot boom; and the TB-2A, a triband beam for $129.95 which features two working elements on 10, 15, and 20 meters. The MB-40H, a new heavy-duty two-element, 40-meter beam is priced at $199.95, and features two working elements on a 15.75-foot steel boom. All Swan beam antennas are rated for 2000 watts PEP and are designed for a vswr of 1.5:1 or better at resonance.

The deluxe mobile models include a five-band mobile 45 antenna which features all band manual switching



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for 10, 15, 20, 40 and 75 meters, a *High-Q* tapped coil, eight positive stop manual positions with gold-plated contacts, featuring a base section, mobile coil and 6-foot whip top section. It is power rated at 2000 watts PEP; cost is $119.95.

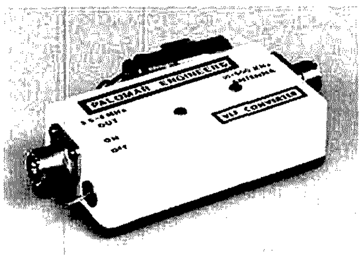
The new Swan 742 tri-band antenna, priced at $109.95, which, once adjusted to desired operating frequency for 20, 40, and 75 meters, requires no further adjustment. It is power rated at 500 watts PEP.

The Golden Swan Trap Vertical antenna, Model 1040V, an omnidirectional, low radiation angle unit designed for 52 ohm coaxial feedline is priced at $122.95. Power rated at 2000 watts PEP, it measures 21 feet high and covers 20, 15, 20, and 40 meters. A 75-meter add-on kit is available for $39.95.

Accessories include a *Kwik-on* connector for easy installation and removal of the mobile antenna for $7.95; an MMBX mobile impedance in-line, low loss match box, $23.95; and WM 3000 in-line peak reading wattmeter for $79.95.

For further information about products and prices, contact Swan Electronics, 305 Airport Road, Oceanside, California 92054.

## 10-500 kHz vlf converter




Palomar Engineers has introduced a new vlf converter which converts signals in the 10-500 kHz vlf band to the amateur 80-meter band so they can be heard on an ordinary shortwave receiver. The converter provides reception of the 1750-meter band at 160-190 kHz where transmitters of one-watt power can be operated without FCC license. It also covers the navigation radio-beacon

band, standard frequency broadcasts, ship-to-shore communications, long-range navy transmitters, and the European low-frequency broadcast band.

The Palomar converter is simple to use and has no tuning adjustments. Tuning of vlf signals is done by the associated receiver which picks up 10-kHz signals at 3510 kHz, 100-kHz signals at 3600 kHz, and 500-kHz signals at 4000 kHz. The converter features crystal control for accurate frequency conversion, a low noise rf amplifier for high sensitivity, and a multi-pole filter to reject broadcast and shortwave interference. Price is $55.00 postpaid in the United States and Canada. For more information write to Palomar Engineers, Post Office Box 455, Escondido, California 92025.

## co-resident 8080 editor/assembler

Tychon has announced its co-resident editor/assembler (TEA) for 8080 systems. Requiring only 5k of memory (R/W or PROM) it is completely I/O independent and relies upon its own I/O software or the I/O routines already available in a user's system. The Tychon Editor/Assembler accepts both octal and hexadecimal values throughout the program and the program listings may be in either octal or hexadecimal form. The switch between octal and hex is made at any time using keyboard commands. The TEA package is the only editor/assembler available which allows the user to easily change the numbering system used. The editor/assembler is relocatable using a special relocator within the program which will place TEA anywhere in the 8080's memory space. The program is available in 1702A or 2708 PROMs and on paper tape. Listings are also available. Prices start at $35 for a paper tape version plus the User's Manual. For further information contact Tychon, Inc., Blacksburg, Virginia 24060.



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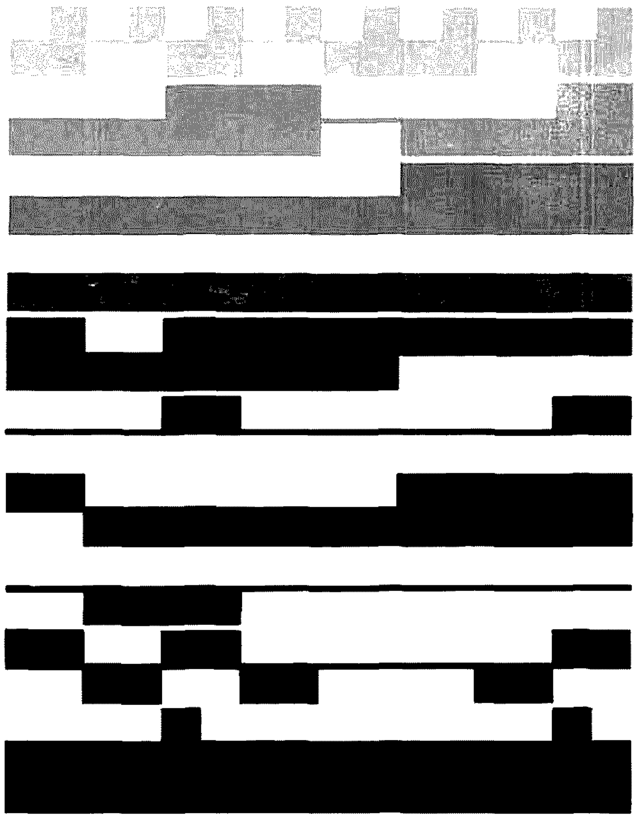
$1.50

# ham radio

magazine

## SEPTEMBER 1977





interlaced
sync generator
for ATV camera control

# ham radio magazine

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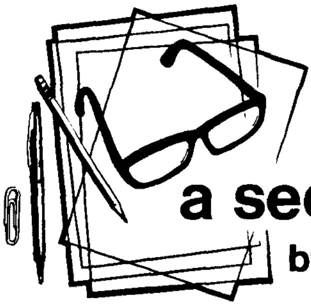


# a second look

by Jim Fisk

At a recent meeting with FCC staffers in Washington, it wasn't surprising that the two items which received the most attention were type acceptance and the proposed linear amplifier ban. Both were brought on by the illegal use of amateur equipment by CBers, and it was generally agreed that neither type acceptance nor an outright ban on linear amplifiers would cure the basic problem — that can be solved only by preventing amateur equipment from getting into the hands of unlicensed operators.

The proposal by the San Antonio Repeater Organization (SARO) that would require presentation of a valid amateur license at the point of sale before amateur equipment could be purchased (*Second Look*, June) is one possible answer to the problem which has been widely endorsed by the amateur equipment manufacturers. Unfortunately, members of the FCC legal staff don't feel the FCC has the authority to impose such a regulation under their present charter. The Communications Act is presently being overhauled by Congress, however, and it is suggested that amateurs write to their Congressmen, asking that authority for point-of-sale control be given to the FCC.

The recently formed Amateur Radio Manufacturers Association (ARMA) has endorsed the basic SARO proposal, as have Dentron, Drake, Heath, Kenwood, and most other large manufacturers. Only one major amateur equipment manufacturer has refused to endorse point-of-sale control, and it's widely reported that their transceiver sales to CBers are greater than those to the legitimate amateur market! Many of the amateur manufacturers and dealers have also agreed to follow ARMA's guidelines for point-of-sale control, but a voluntary program is not likely to be very successful; if a CBer is unable to buy the amateur equipment he wants from his local dealer, both Sears and Wards list amateur equipment in their catalogs, and neither, apparently, has any inclination to require the purchaser to produce a valid amateur license. So while the flow of amateur equipment into CB hands can be slowed down, without the necessary rule making by the FCC, it can't be completely stopped.

As further proof that the ban on 24- to 35-MHz linear amplifiers is unworkable, at least one resourceful manufacturer is now selling "amateur" six-meter linears — the user simply has to remove a jumper across the tuning coil to put the unit on 27 MHz!

Here at *ham radio* we have been occasionally faced with the problem of deciding what is, and what obviously is not, a valid amateur product. More than one manufacturer of an "amateur" linear has tried to legitimize his product by advertising the unit in an amateur magazine. Since we had no published guidelines on the subject, in at least one case we were forced, under threat of legal action, to accept advertising for a product which we felt had no place in *ham radio*. Therefore, we have advised all equipment manufacturers that advertising for external rf power amplifiers designed for use on frequencies below 60 MHz, other than those operating class C, must meet the following requirements:

1. Amplifiers must meet all applicable FCC requirements for operating in the Amateur Radio Service.
2. Amplifiers shall be capable of operating with at least 50 watts rms rf input drive power without exceeding FCC specifications for spurious and harmonic output.
3. Amplifiers shall be bandswitching and shall have no 11-meter (27 MHz) bandswitch position.
4. Amplifiers shall require the use of an *external* ON-OFF keying line for transfer from the standby to operate modes of operation.

These standards have been designed to reflect the requirements and practices currently in use in the Amateur Radio Service. As the state of the art advances, it may be necessary to revise these requirements, but there will be no special exceptions or case-by-case dispensations. This policy becomes effective with the October, 1977, issue of *ham radio*.

There is probably no way to completely cut off the flow of amateur equipment into the CB marketplace, but if it can be reduced to a trickle, perhaps amateurs won't be saddled with regulations which would virtually eliminate commercial linears that operate on the amateur 10-meter band.

**James Fisk, W1HR**
**editor-in-chief**

# presstop

COMMUNICATOR LICENSE WAS KNOCKED DOWN but not entirely out by the FCC in late July. As proposed, the Communicator privileges would have been phone only on 220-225 and 420-450 MHz, reserving 435-438 for satellite communications. The Communicator would become the entry level license, with element 2 (the present Novice written exam) re-oriented to include phone, and Field Office administered. To upgrade to Novice, a Communicator would pass a volunteer-administered CW exam.

The Commissioners' Final Vote was unanimous against funding the Communicator in next year's budget. Their rejection was tempered, however, by a recommendation that the concept be recoordinated with the objectors and then resubmitted at a later date. Rejection at this time pretty well pushes the time-table for the Communicator back another year, until early 1980.

FCC-HOSTED "MEDIA WORKSHOP" in Washington July 13th provided an FCC/Amateur rap session that was rated "simply outstanding" by the 50 or so who attended. The well-filled and well-organized program lasted from 9:00 AM until almost 5:00, with lunch and a short break to see the Commissioners at work on several Amateur agenda items (they okayed "AAx2" callsigns for the Extra Class callsign program) the only pauses. As an open, public meeting the limits placed on Amateur-FCC dialogue by the Home Box Office rule did not apply. Personal Radio Division Chief John Johnston chaired the session smoothly but flexibly, with give-and-take rather than formal type presentations the format. Only limitations were that we were not to "advise" or "recommend" — only "discuss" and "provide information."

ARRL'S BOARD MEETING in Hartford July 21-22 covered a mixed bag of topics, old and new, recording some very significant accomplishments. One of the more important decisions was to have the League become more active on the Washington scene, with a Newington staffer to spend full time on ARRL Washington activities; the League President, Vice President, General Manager, and General Council designated to represent the ARRL at forthcoming Congressional hearings that concern Amateur Radio; and the taking of necessary steps to ensure that the League complies with Lobbying regulations while maintaining an effective voice in Washington.

RUSSIAN AMATEUR SATELLITE'S FREQUENCIES have just been officially filed by the Russians with the International Frequency Registration Board, further confirming the impending launch of a Soviet Amateur spacecraft. Called the "USSR Amateur System 'RS" in the IFRB filing, the spacecraft is to have a 950 km (590 mile) high gear circular, 102 minute orbit with 82$^{o}$ inclination, with a 145.8-145.9 MHz in, 29.3-29.4 MHz out, 1.5 watt peak output transponder. Three to four satellites in all are proposed, to be launched in 1977-1978.

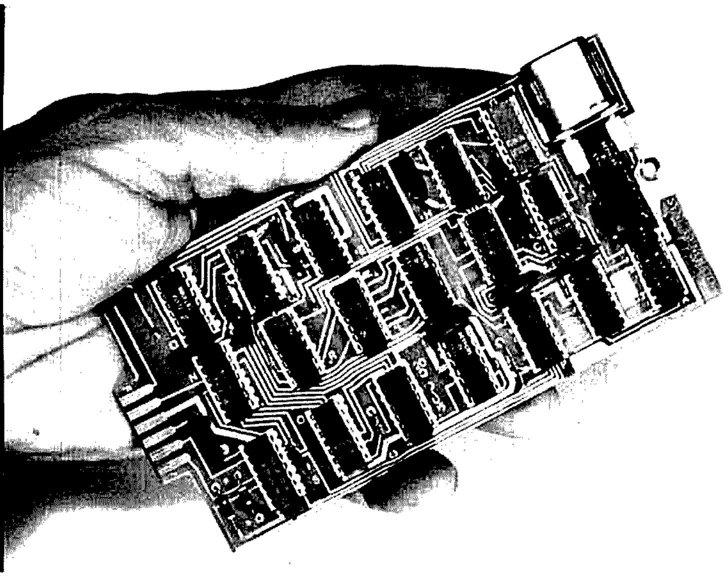
AMSAT's AO-D Launch is now firming up for February 23 and is planned to put the satellite into 102.8 minute orbit with an apogee of 935.4 km (577.38 miles) and a perigee of 888.8 km (548.665 miles).

AMATEUR LICENSE FORMS have arrived at Gettysburg to break up an almost six-week logjam in Amateur license distribution. An expedited preliminary shipment of 50,000 forms showed up in early August, and FCC personnel were hopeful they'd have the oldest of the backlogged licenses printed and in the mail within a week.

TWO-METER SIGNALS CROSSED THE ATLANTIC the end of June when PY2OB in Sao Paulo, Brazil heard TU2EF in the Ivory Coast on 145.2-MHz CW! A two-way contact didn't result at the time, but a series of followup attempts were under way by TU2EF and TU2GA with PY2OB and other Brazilian VHF buffs on the other end of the 3496-mile (5665km) circuit.

On Six Meters, long haul has also been prevalent with a contact between WB2RLK/VE1 and KH6HI, and a number of U.S. stations working into northern South America. The 50-MHz beacon on Gibraltar is supposed to be back on the air shortly, supplementing France's FX3VHF on 50.1 MHz while South Africa has another form of six-meter beacon with a new channel 1 TV station.

HAM RADIO/HAM RADIO HORIZONS Assistant Advertising Manager, Cindy Schlosser, left Greenville August 12th to become Advertising Manager for Solar Age, a trade magazine. A nice upward move for Cindy, who'll be missed in Amateur circles.



# interlaced sync generator

## for ATV camera control

Complete, interlaced camera control is provided by this digital sync generator

**Over the last few years** a number of fast-scan television sync generators have been described, some simple and others complex. In general, most of these designs were either too costly or they required expensive test equipment to accurately adjust pulse widths and timing. Recently though, a number of manufacturers have developed single IC sync generators. However, both the cost of the IC and the amount of additional logic required makes that approach rather expensive. (Although it does eliminate the need for external setup adjustments). This sync generator provides all of the control signals that are required to run a broadcast-quality television camera, yet it is inexpensive and requires no adjustments to provide accurate pulse timing.

I believe the sync generator to be described here eliminates both high cost and external setup drawbacks by providing the following features:

1. Readily available TTL ICs are used throughout, providing easy access to parts.
2. All components are inexpensive.
3. No timing or alignment adjustments are required.
4. Low power requirements simplify power supply design, + 5 volts at 350 mA.
5. A free-running oscillator can be used for those designs not needing extreme frequency stability.
6. Compact size. The entire circuit (exclusive of the power supply) can be built on one 3 x 6 inch (7.6 x 15.2cm) circuit board.

This sync generator has a number of very practical

**By Arthur Towslee, WA8RMC,** 180 Fairdale Avenue, Westerville, Ohio 43081

uses considering its cost and overall size. My primary purpose is for ATV interlaced camera control. A future article will describe full construction details of the RMC (**R**eliable **M**ini **C**amera). Other applications include main timing for an accurate bar-dot generator for TV alignment or the master control for TV character generators.

## general description

A block diagram of the unit is shown in fig. 1. All

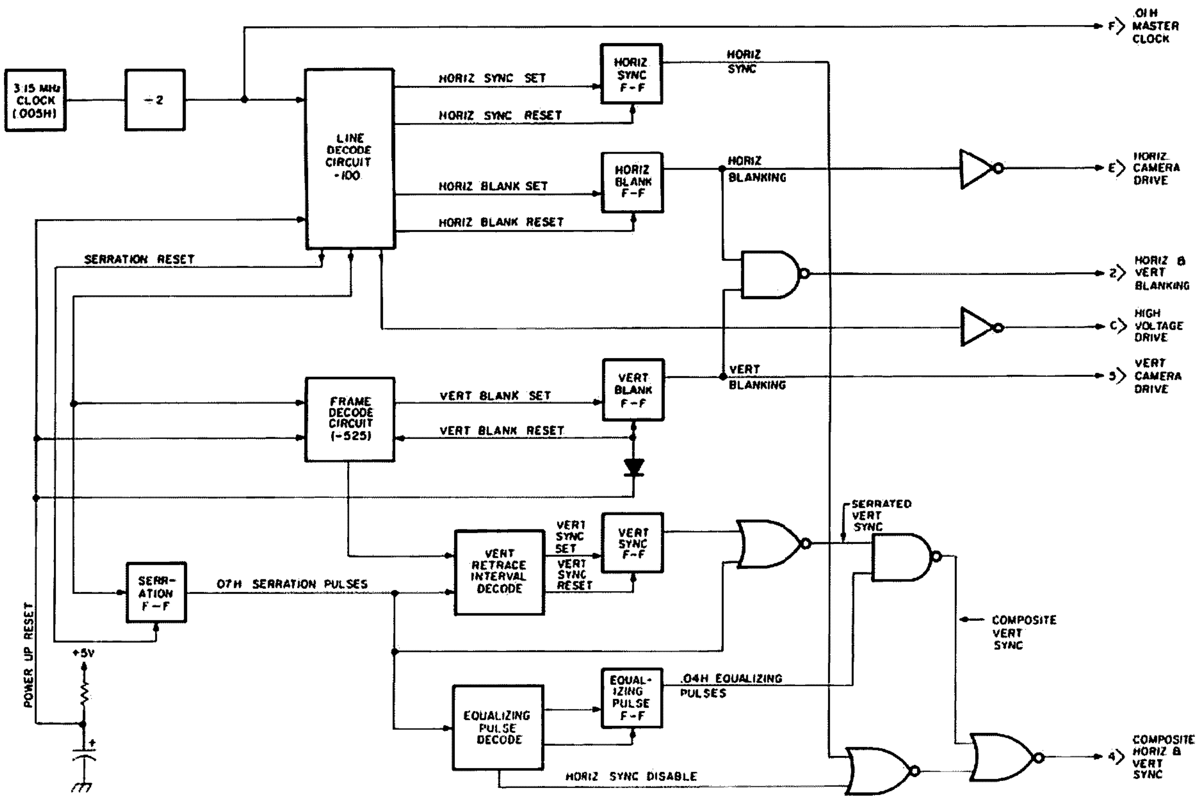



**fig. 1. Block diagram of the complete interlaced sync generator.**

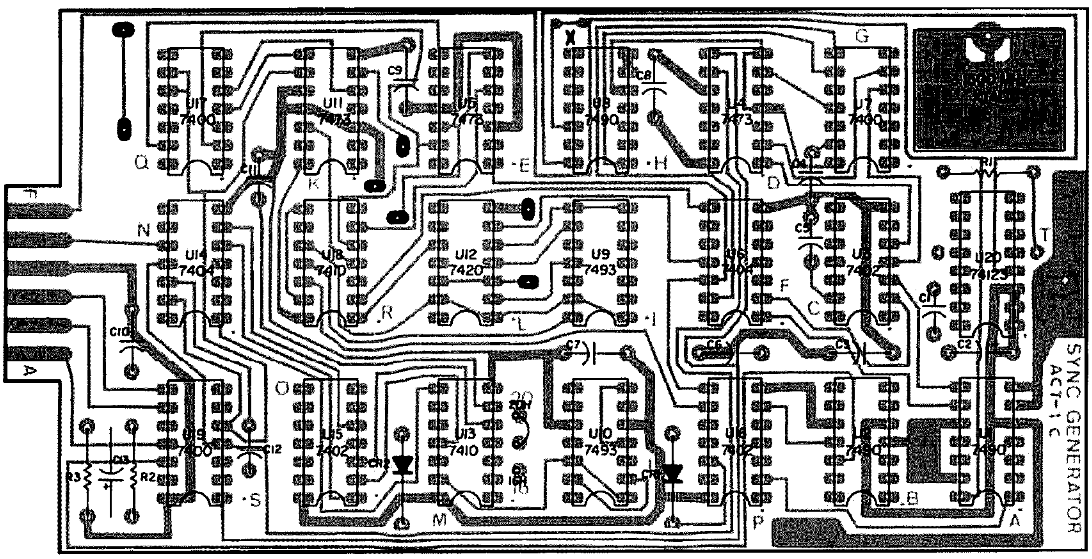



**fig. 2. Parts overlay for the generator. The oval pads indicate a feedthrough point which does not contain a component. The jumper must be connected for the desired vertical blanking time. Pin 1 of each IC is designated on the circuit board to prevent installing the IC improperly. The crystal socket is an Augat 8000D or equivalent.**

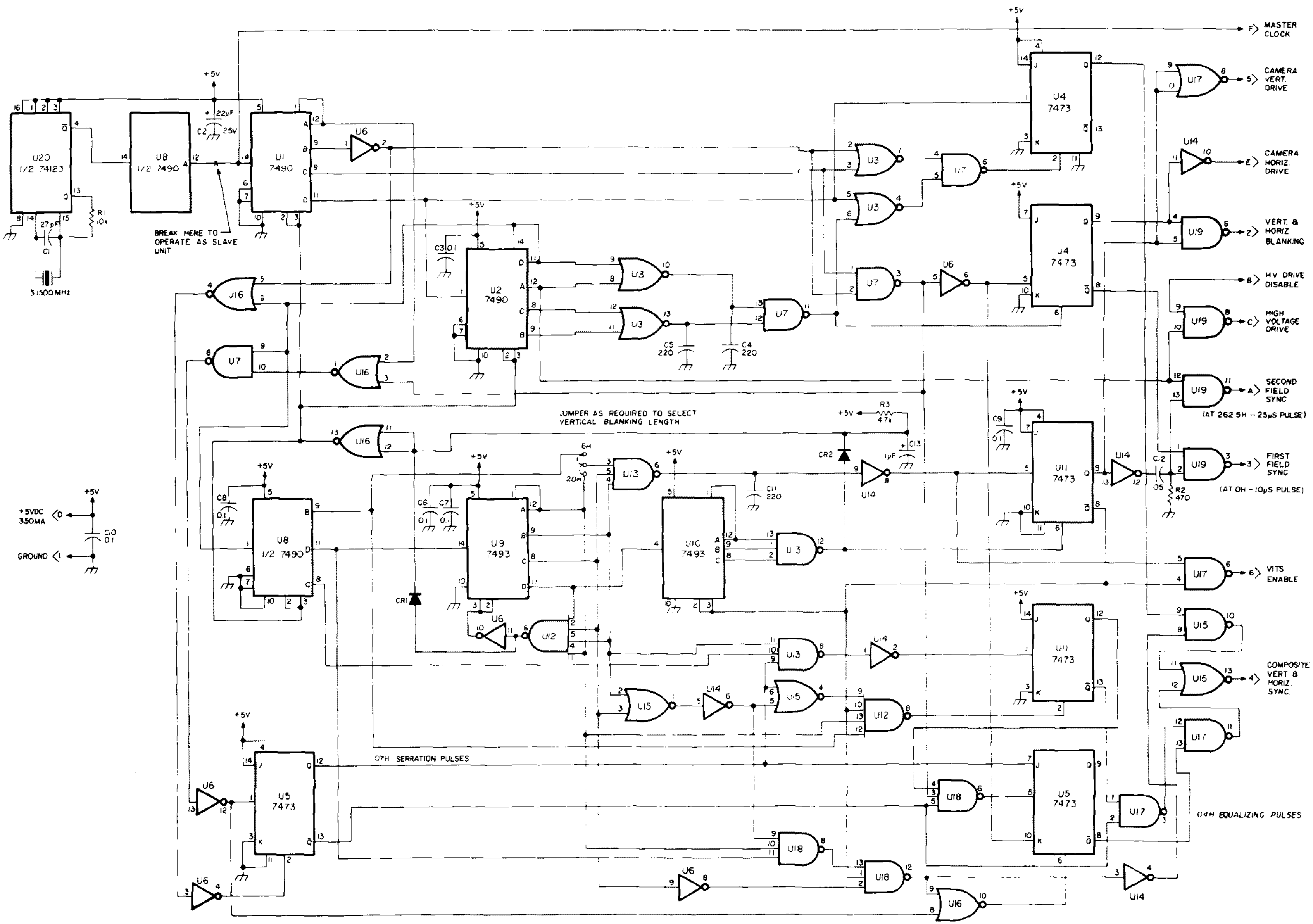



fig. 3. Schematic diagram of the sync generator. Capacitors C3 through C12 are disc ceramics; resistors R2 and R3 are ¼-watt. CR1 and CR2 (1N270 or 1N47A) and C13 are omitted when slaves are not used.

ICs are standard, easy-to-obtain 7400 series TTL devices mounted on a single 3 x 6 inch (7.6 x 15.2cm) double-sided printed circuit board. All outputs are standard totem pole voltage levels. These outputs will be described using the following definitions: H is the total time required for the scanning beam to travel across one horizontal line and is equal to 63.5 microseconds. V is the time required for the same beam to travel from the top left to the bottom center of the TV raster and is equal to 262.5 horizontal lines or 16.66 milleseconds. V is equal to one picture field where two fields are required for each frame. One complete frame contains 525 horizontal scan lines. Therefore, 2V = 33.33 milliseconds or 1/30 Hz, producing 30 complete frames per second.

## outputs

Before discussing the outputs (**fig. 3**), there should be an understanding of the terminology used. When referring to the logic signals, either plus or ground is used to denote the true state. For example, if the desired signal is a positive-going pulse it would be referred to as a plus true signal. For a negative pulse, it would be ground true.

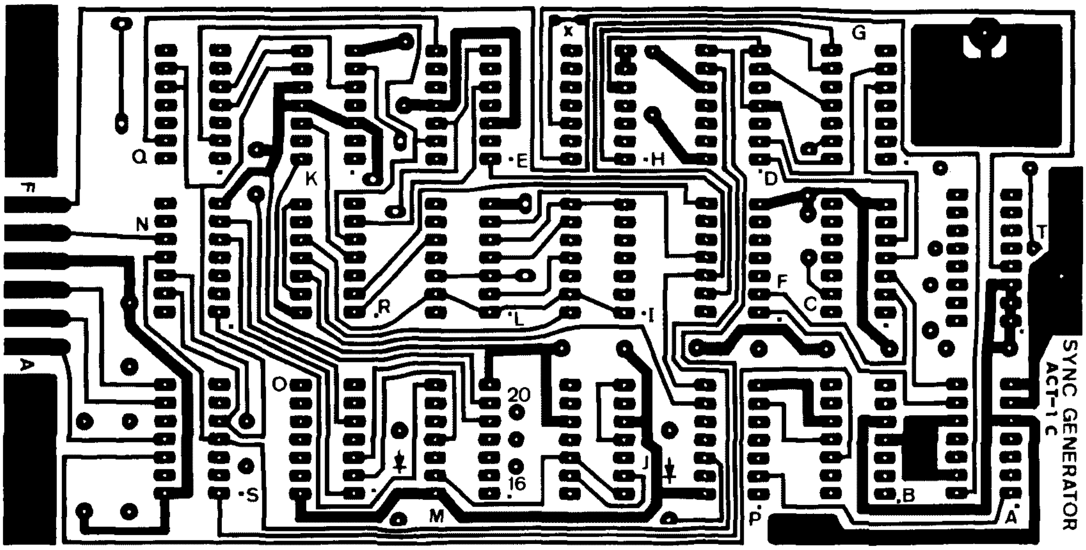



**fig. 4. Circuit board layout for the top side of the board. At the point on the board marked with a small X, the trace can be broken for slave operation as explained in the text.**

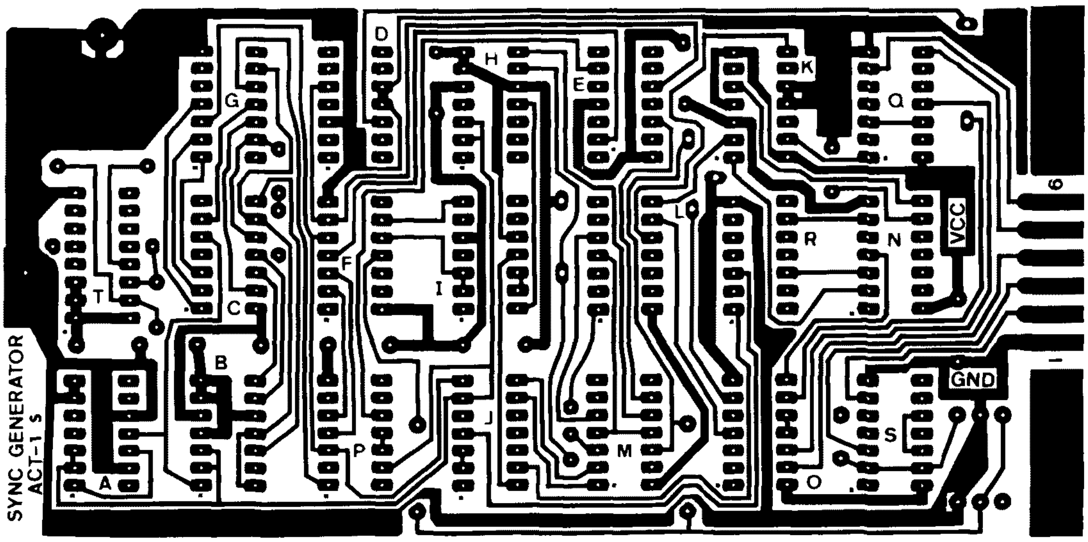



**fig. 5. Layout for the bottom side of the printed circuit board.**

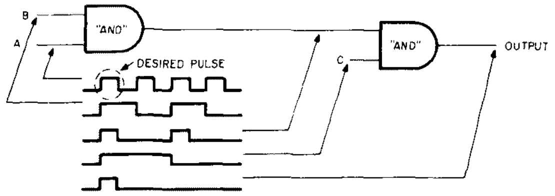



**fig. 6. Method of windowing pulse trains to obtain a single desired pulse.**

**1.** Pin E — Camera horizontal drive. This is a plus true pulse 0.16H wide, starting at zero H, and is used for driving the camera horizontal sweep circuit and for vidicon horizontal blanking.

**2.** Pin 5 — Camera vertical drive is a plus true pulse 20H wide starting at zero V and is used for driving the camera vertical sweep circuit and for vidicon vertical blanking.

**3.** Pin 2 — Processor mixed blanking. This is a plus true pulse that is the combination of the vertical and horizontal drive pulses. It is used in the video processor to insert blanking into the video.

**4.** Pin 4 — Processor composite sync is a combination of ground true pulses composed of serration pulses, equalizing pulses, vertical sync, and horizontal sync pulses. A more detailed description of this pulse chain will be covered later.

**5.** Pin F — Master clock. This pulse chain, with a symmetrical period of 0.01H, is used when it is desired to slave one or more sync generators from a master unit. On the slave units, point **X** is broken and the crystal removed.

**6.** Pin C — High voltage drive. This is a symmetrical square wave signal with a period of 1H used to trigger a high-voltage converter for vidicon operating voltages.

**7.** Pin B — High voltage disable is a ground true *input* that will kill the high-voltage drive upon the failure of any correct operating condition. If this pin is not used, it must be connected to pin D ( + 5V).

**8.** Pin 3 — First field sync. This is a ground true pulse approximately 10 microseconds wide starting at zero H. It facilitates troubleshooting the circuitry by providing an oscilloscope trigger to view the first (odd) field.

**9.** Pin A — Second field sync is a ground true pulse approximately 25 microseconds wide starting at 262.5H. It provides triggering to view the second (even) field.

**10.** Pin G — VITS enable. This pulse is ground true, 2.5H wide starting at 17.5H and 280H. It is used to enable test signals presented to the video waveform for 2.5 horizontal lines immediately prior to unblanking of the waveform. Broadcast TV uses this time to generate **V**ertical **I**nterval **T**est **S**ignals critical to quality analysis of the video waveform. However, little if any use of this feature is needed in most television cameras.

**11.** Pin 1 — Ground

**12.** Pin D — Vcc, + 5 volts at approximately 350 mA.

**Master oscillator.** The master oscillator is a novel design[1] using half of a 74123 dual one-shot retriggerable multivibrator. A 74122 may be used instead, but it is not pin compatible. I believe the 74123 is also a bit easier to obtain. Minimal parts are needed and the output is a symmetrical TTL compatible square wave. R1 and C1 form a series resonant RC circuit slightly above the crystal frequency. Without the crystal the oscillator would free run at the frequency determined by these values. The 74123 **Q** output starts in a low state and switches to a high state after C1 charges. The high state is unstable, however, and the one-shot discharges the **Q** output through R1 returning **Q** to a low state which repeats the cycle. Even though I use a crystal to make this oscillator stable, by careful selection of R1, a good metal film resistor, and C1, a good silver mica capacitor, the oscillator will free run with sufficient stability

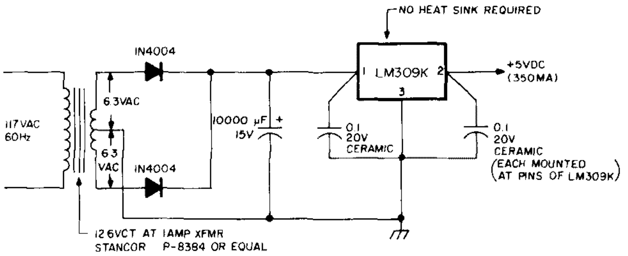



**fig. 7. Typical power supply for the interlaced sync generator.**

for ATV operation. I use an International Crystal type Ex crystal because of its low cost and easy crystal availability.

Following the oscillator is a simple divide-by-two flip-flop to produce the main control clock for the rest of the circuitry. The total period at this point is 0.01H which has a low state for 0.005H and a high state for 0.005H. At this point the circuit may be broken for slave operation from another source or it could serve as the master clock for other devices.

**Line decoding.** A number of counters and flip-flops combine to form *window* circuits for the purpose of extracting the horizontal sync and blanking pulses (**fig. 8**). A total frequency division of 100 provides a

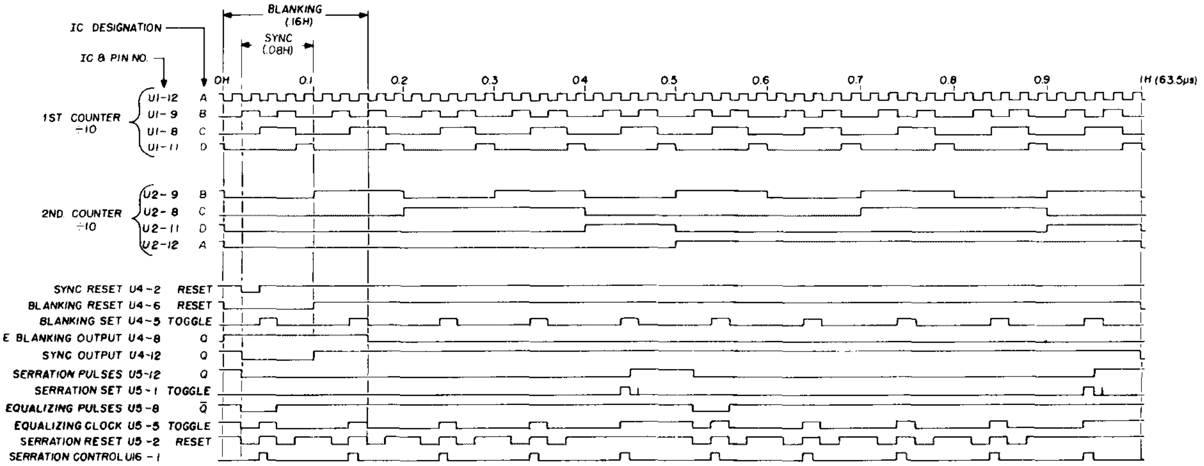



**fig. 8. Horizontal interval waveforms. Note that the sync pulse starts after blanking starts and ends before the picture is unblanked.**

1H pulse chain to drive the vertical circuits. The windowed pulses from U1 and U2 combine to form the set and reset lines for the J-K flip-flops, U4.

It is appropriate at this time to briefly describe window circuits because the entire logic is built upon this approach. To window a pulse chain simply means to extract only a portion of the pulse information, with respect to time. For example, in **fig. 6**, a particular line (input A) has four pulses in a given time frame and it is desired to extract only the first one. This series of pulses is logically ANDed with pulses of a lower repetition rate (input B), producing an output only when both inputs are plus. Therefore, of the four pulses at input A and the two pulses at input B, the output is true for only two of them. This is subsequently repeated with a third pulse train (input C) to produce the desired output. By extending this approach, it is easy to identify one pulse in a series of thousands, within any given time period.

**Frame decoding.**, The 1V time interval requires 1H to be divided by 525. The combination of U8, U9, and U10 provide division by 5, 15, and 7, respectively, for a total of 525. The outputs then feed U11 to obtain the basic vertical frame rate (**fig. 9**). I've provided an optional vertical blanking time of either 16H or 20H, the latter being the standard. If it is desired to unblank after the 16th horizontal line, providing four extra lines of active video, put the jumper in the corresponding position. In reality, this feature has only minimal value and, I confess, wouldn't have been provided had it not been for an available 3-input gate with just 2 inputs used.

**Vertical retrace and equalizing pulse decoding.**

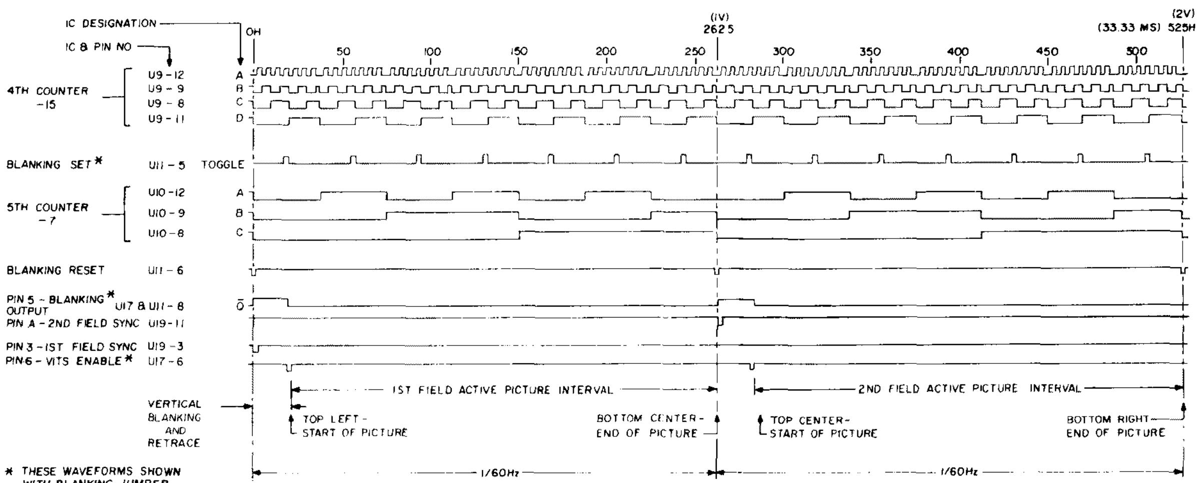



**fig. 9. Vertical interval waveforms. Both fields are shown. The first field starts at OH and the second at 262.5H. The first field covers from the top left to the bottom center of the picture while the second covers the bottom center to bottom right.**

This circuitry produces the vertical serration, sync, and equalizing within the vertical blanking time. No further frequency division is required, but some rather fancy manipulation of flip-flops and window circuitry is necessary to generate the waveforms required by EIA standards. The serration pulses are combined with the vertical sync to produce serrated vertical sync (U17 p3) which is then added to the equalizing pulses, providing composite vertical sync serration pulses determine proper operation of the vertical retrace and also the accuracy of the interlace. Good interlace is not only determined by the quality of the camera, but also by the proper functioning of the receiver sync separator. Providing a crystal-controlled time base and dividing the horizontal frequency by 525 by itself does *not* guarantee true interlace!

**Logic reset circuit.** Although no power up resetting

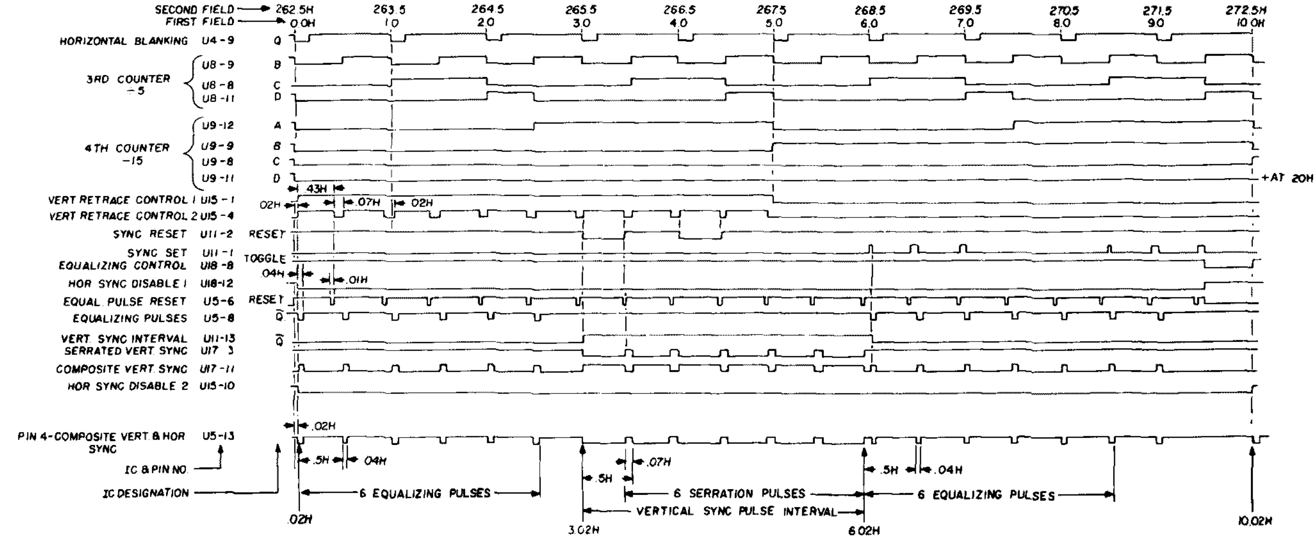



**fig. 10. Vertical retrace interval waveforms. This illustrates how the composite sync is constructed.**

(U17 p11). The horizontal sync is next added to produce composite processor sync which contains serrations, equilizing pulses, vertical sync, and horizontal sync. A basic understanding of the significance of serration and equalizing pulses is required to fully appreciate the required logic.

At every 1H, or multiple thereof (**fig. 10**), a sync pulse must occur to keep the receiver horizontal oscillator on frequency. Therefore, equalizing and serration pulses occur at a 1/2H repetition rate during vertical blanking to provide a uniform transition from the even to the odd fields. Also, looking at the receiver's sync separator circuitry, it's observed that the capacitor-coupled integrator depends upon average dc level to separate vertical from horizontal sync. Therefore, the negative-going equalizing pulses help establish this transition by providing a higher positive average voltage just prior to vertical sync. When the transition to vertical sync takes place, the positive-going serration pulses keep the horizontal oscillator locked while producing a lower average dc level. The receiver integrator charges at this point until it resets its own vertical ramp and retrace takes place, so the existence of equalizing and

is required for normal single camera operation, provisions have been made when using a master-slave approach to reset all generators to zero and start from the same point. This can be accomplished by making the value of C13 significantly larger in the master unit than in each slave. Then at power up, all units will be reset to zero and held there for a time determined by C13 and R3. If the master unit has a longer time constant it will be held reset longer than each of its slaves. Since the master feeds the control clock to each slave, when the master is reset the slaves also stay reset. The purpose is to prevent the master unit from feeding pulses to the slaves before they complete the reset command. If the generator is not used with slaves, C13 as well as CR1 and CR2 may be omitted.

One final point, the placement of CR1 and CR2 appears to violate the loading rules of TTL active gate outputs. However, the time duration that a short circuit exists is not long enough to present

*A drilled printed circuit board for $9.00 postpaid is available from Automation Engineering Company, 3621 Marine Drive, Toledo, Ohio 43609.

problems, and is within the safe operating parameters of these devices.

The complete sync generator is contained on a 3 x 6 inch (7.6x15.2cm) *double-sided printed-circuit* board. Artwork for the top and bottom is shown in **figs. 4** and **5** for those who wish to duplicate the board.* If desired, breadboard or wire-wrapped versions can be constructed if the following precautions are observed. First, all leads should be as short as possible. This is especially true in the oscillator and counter (U1) areas. The high frequencies involved could cause noticeable radiation to other units, especially video preamps.

Make liberal use of bypass capacitors at the +5 volt terminal of each counter or flip-flop. A good ceramic capacitor of at least 0.1 $\mu$F or higher should be used. Note also that I have used 220 pF capacitors at strategic points in the circuit. This is to eliminate possible switching spikes being generated in the ripple counters. It is possible to redesign this circuit using synchronous counters, but no problems have been encountered in a number of units constructed. Finally, I do not recommend sockets for the integrated circuits! More time has been spent troubleshooting faulty socket connections than for bad ICs.

## power supply

As mentioned earlier, a single 5 volt, 350 mA regulated power supply is required. A simple rectifier circuit followed by an LM309K regulator (**fig. 7**) is quite satisfactory. The regulator does not require a heatsink.

## conclusion

A good sync generator, easy on the pocketbook and also to build, is always welcomed by the serious ATV enthusiast. I believe that this article meets these criteria. My intentions were not to dive into a theory session on TV sync generation, but to accentuate some key points that have been left out of previous articles on amateur television. A future article will describe a complete television camera using four printed-circuit boards including this sync generator.

## references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2. Gerald A. Eastman, "Measurement Concepts," Tektronix, Incorporated, 1969, page 44.
3. "Television Signal Analysis," Second Edition, American Telephone & Telegraph Company, April, 1963, page 5.


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1. Mark Trueblood, "Three Components Make A Stable Crystal Oscillator," *EDN Magazine*, July, 15, 1972, page 54.


**ham radio**

# tracking the OSCAR satellites

## Orbital positions and your range capabilities can be determined easily with the aid of a few simple charts

**Many articles**, aids, information, and data[1-6] have been published on the use of the OSCAR 7 satellite. However, some questions still remain unresolved: How do you work over long distances? How do you communicate with the satellite over periods of one or two minutes? How do you establish the physical limitations of your communications capabilities?

The article attempts to provide answers to these questions. All you need are the few easy-to-make visual aids described here and an understanding of some of the basic concepts about OSCAR 7 and your ground station.

### background

OSCAR 7 is located 900 nautical miles (n. mi.) (1036 statute miles or 1668 km) above the earth's surface.* OSCAR 7 completes one orbit around the earth every 114.9 minutes. The satellite moves at about 23,000 fps (7000 mps). It can see a section of the earth's surface of 2250 n. mi. (4167 km) in all directions.

Theoretically, if you were riding in OSCAR 7, you could see two points on the earth's surface that are 4500 n. mi. (8334 km) apart. There is, though, a very significant limit to OSCAR 7 as a communications tool. OSCAR 7 is usable only if it's in the visibility region of your ground station. Your visibility region is limited to the line-of-sight distance between your station and the satellite as it moves in space. Unless two stations on the earth's surface are closer than 4500 n. mi. (8334 km) they can't communicate, since OSCAR 7 is a line-of-sight repeater.

Each operator using OSCAR 7 is faced with the following problems. He must be able to **1)** locate OSCAR 7, **2)** establish what cities or stations are within his maximum communications capability, and **3)** know when and where OSCAR 7 will be passing through the region of mutual visibility.

### graphical aids

These three problems are not difficult to resolve, and learning to use OSCAR 7 is a simple process. The starting place is in constructing or obtaining the few simple graphical aids required. We call them the Precision Orbital Position Plotting Chart, (POPP chart is shown). The POPP has been instrumental for establishing satellite communications between San Jose, California and Sapporo, Japan during communications periods as short as one minute. The POPP chart is shown in two forms: a polar-sterographic map projection of the northern (or southern) hemisphere, **fig. 1**, and a Mercator map projection of

*Distance measurements are given in nautical miles (n. mi.), which are longer than the statute mile by 1.15. The nautical mile is used for navigational purposes. Each nautical mile is the distance of longitude measured at the equator (6086 feet or 1.85 km). **Editor.**

**By W. R. Harmon and N. Patrick Peterson, WA6UAP.** Mr. Harmon's address is 10560 Stokes Avenue, Cupertino, California 95014; WA6UAP may be reached at 1422 Bretmoor Way, San Jose, California 95129

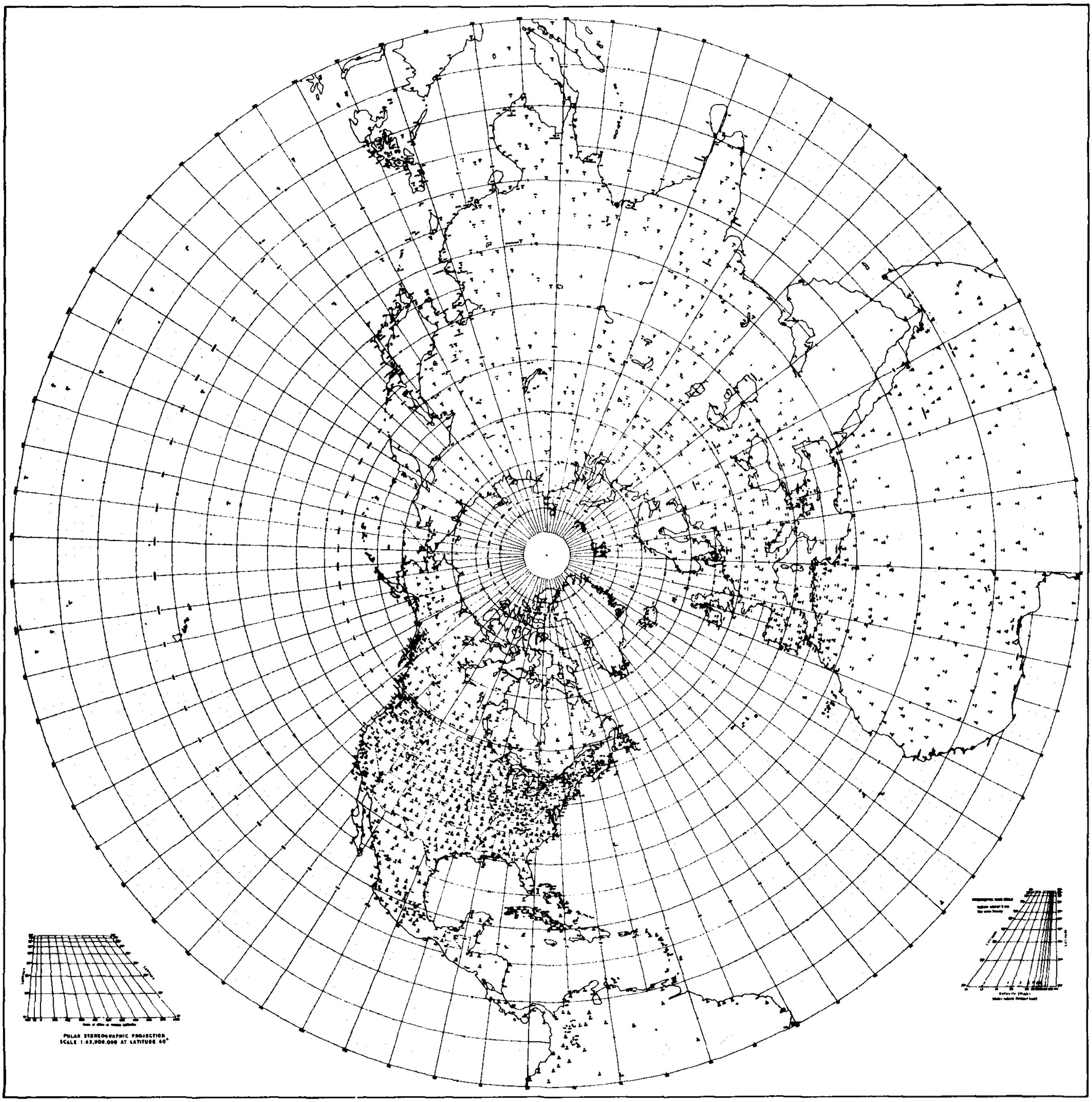


fig. 1. Polar-stereographic projection of the northern hemisphere used for the Precision Orbital Position Plotting (POPP) chart. A similar projection for the southern hemisphere is available from *ham radio*.*

the world, fig. 2. The polar-stereographic projection is preferred for good reason; because of the geometric characteristics of such a projection, the ground-station mutual visibility regions may be plot-

*Large polar-stereographic projections of the northern and southern hemispheres, suitable for constructing your own graphical tracking aid, are available for $1.00 postpaid from *ham radio*, Greenville, New Hampshire 03048.

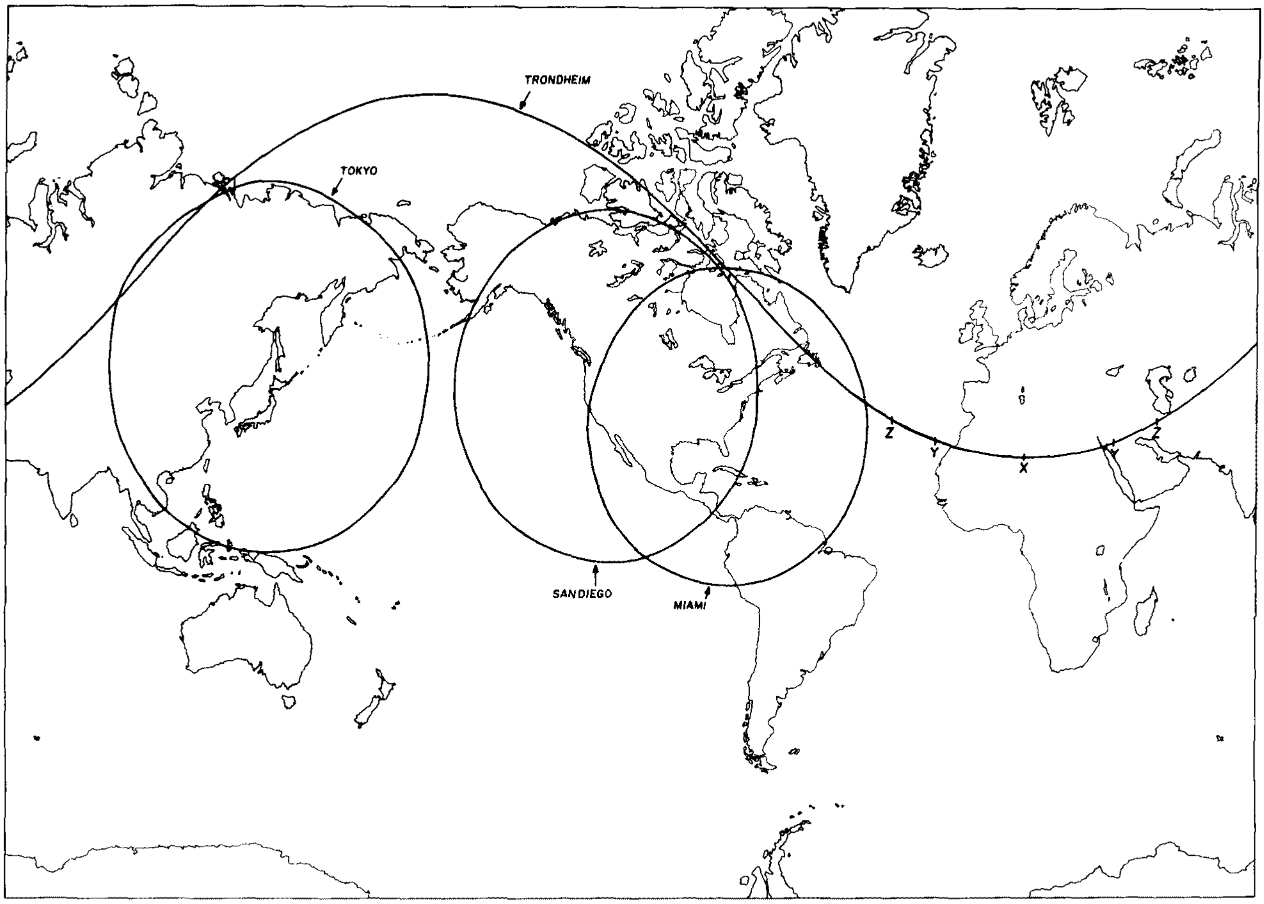



**fig. 2. Mercator projection of the world showing visibility regions of four ground stations. Note that the visibility regions are not circles. This projection is less desirable for a POPP chart than the polar-stereographic projection of figs. 1 and 3.**

ted as perfect circles, **fig. 3**. This is not true for the Mercator projection, nor for the polar projections which are usually used by Amateurs for satellite tracking. The Mercator projection is complex and difficult to use, particularly for stations at latitudes more than 45 degrees north or south.

The steps in making a POPP chart for use with OSCAR 7 are:

**1.** Inscribe on your selected map projection the 2250 n. mi. (4167 km) visibility region of your ground station. Also mark the 4500 n. mi. (8334 km) range from your station.

**2.** Obtain an orbit ground track for a single OSCAR 7 pass and plot this orbit on a clear acetate overlay of the map you plan to use. The orbital path should include the times marked at one-minute intervals (see **figs. 4** and **5**).

**3.** Obtain the schedule of the OSCAR 7 longitudes of ascending node (LAN) times and locations.[7,8,9] The LAN is the time and longitude when the satellite's *ground track* crosses the equator on each south-to-north pass.

**4.** Select a potential contact inside the 4500 n. mi. (8334 km) range and plot his visibility region on your chart. This circle of visibility should include your station visibility region to form a region of mutual visibility.

**5.** With the satellite-orbit overlay, find an LAN that results in the orbit transversing the region of mutual visibility.

**6.** Add the time of the LAN to the elapsed orbit ground-track time entering and leaving the region of mutual visibility. This time period then becomes the *communications-time window*.

**7.** Calculate or estimate your antenna pointing azimuth for the region of mutual visibility.

**8.** Set up equipment, wait for the proper time, then

table 1. Latitude and longitude of OSCAR 7 orbits at one-minute intervals for the northern hemisphere.

| time (min.) | latitude (deg. north) | longitude (deg. west) | time (min.) | latitude (deg. north) | longitude (deg. west) |
|---|---|---|---|---|---|
| 0 | 0 | 140.00 | 29 | 78.23 | 242.37 |
| 1 | 3.07 | 140.89 | 30 | 77.56 | 257.29 |
| 2 | 6.15 | 141.70 | 31 | 76.20 | 270.05 |
| 3 | 9.22 | 142.60 | 32 | 74.33 | 280.25 |
| 4 | 12.29 | 143.60 | 33 | 72.12 | 288.22 |
| 5 | 15.36 | 144.52 | 34 | 69.66 | 294.46 |
| 6 | 18.43 | 145.47 | 35 | 67.04 | 299.42 |
| 7 | 21.49 | 146.47 | 36 | 64.31 | 303.45 |
| 8 | 24.55 | 147.45 | 37 | 61.51 | 306.78 |
| 9 | 27.61 | 148.49 | 38 | 58.64 | 309.60 |
| 10 | 30.67 | 149.50 | 39 | 55.73 | 312.03 |
| 11 | 33.71 | 150.72 | 40 | 52.79 | 314.16 |
| 12 | 36.75 | 151.93 | 41 | 49.82 | 316.04 |
| 13 | 39.79 | 153.21 | 42 | 46.83 | 317.74 |
| 14 | 42.81 | 154.60 | 43 | 43.83 | 319.28 |
| 15 | 45.02 | 156.09 | 44 | 40.81 | 320.70 |
| 16 | 48.81 | 157.74 | 45 | 37.78 | 322.02 |
| 17 | 51.79 | 159.55 | 46 | 34.74 | 323.25 |
| 18 | 54.74 | 161.59 | 47 | 31.70 | 324.41 |
| 19 | 57.66 | 163.91 | 48 | 28.64 | 325.52 |
| 20 | 60.54 | 166.50 | 49 | 25.59 | 326.57 |
| 21 | 63.37 | 169.73 | 50 | 22.53 | 327.59 |
| 22 | 66.13 | 173.50 | 51 | 19.46 | 328.57 |
| 23 | 68.79 | 178.11 | 52 | 16.40 | 329.53 |
| 24 | 71.31 | 183.87 | 53 | 13.33 | 330.46 |
| 25 | 73.62 | 191.20 | 54 | 10.26 | 331.38 |
| 26 | 75.62 | 200.59 | 55 | 7.18 | 332.28 |
| 27 | 77.17 | 212.49 | 56 | 4.11 | 333.18 |
| 28 | 78.09 | 226.79 | 57 | 1.04 | 334.07 |

table 2. Latitude and longitude of OSCAR 7 orbits at one-minute intervals for the southern hemisphere.

| time (min.) | latitude (deg. south) | longitude (deg. west) | time (min.) | latitude (deg. south) | longitude (deg. west) |
|---|---|---|---|---|---|
| 0 | 0 | 334.40 | 29 | 78.23 | 76.77 |
| 1 | 3.07 | 335.29 | 30 | 77.56 | 91.69 |
| 3 | 6.15 | 336.10 | 31 | 76.20 | 104.45 |
| 3 | 9.22 | 337.00 | 32 | 74.33 | 114.65 |
| 4 | 12.29 | 338.00 | 33 | 72.12 | 122.62 |
| 5 | 15.36 | 338.92 | 34 | 69.66 | 128.86 |
| 6 | 18.43 | 339.87 | 35 | 64.31 | 137.85 |
| 7 | 21.49 | 340.87 | 36 | 64.31 | 137.85 |
| 8 | 24.55 | 341.95 | 37 | 61.51 | 141.18 |
| 9 | 27.61 | 342.99 | 38 | 58.64 | 144.06 |
| 10 | 30.67 | 343.90 | 39 | 55.73 | 146.43 |
| 11 | 33.71 | 345.12 | 40 | 52.79 | 148.56 |
| 12 | 36.75 | 346.33 | 41 | 49.82 | 150.44 |
| 13 | 39.79 | 347.61 | 42 | 46.83 | 152.14 |
| 14 | 42.81 | 349.00 | 43 | 43.83 | 153.68 |
| 15 | 45.02 | 350.49 | 44 | 40.81 | 155.10 |
| 16 | 48.81 | 352.14 | 45 | 37.78 | 156.42 |
| 17 | 51.79 | 353.95 | 46 | 34.74 | 157.65 |
| 18 | 54.74 | 355.95 | 47 | 31.70 | 158.81 |
| 19 | 57.66 | 358.31 | 48 | 28.64 | 159.92 |
| 20 | 60.54 | 0.90 | 49 | 25.59 | 160.97 |
| 21 | 63.37 | 4.13 | 50 | 22.53 | 161.99 |
| 22 | 66.13 | 7.90 | 51 | 19.46 | 162.97 |
| 23 | 68.79 | 12.51 | 52 | 16.40 | 163.93 |
| 24 | 71.31 | 18.27 | 53 | 13.33 | 164.86 |
| 25 | 73.62 | 25.60 | 54 | 10.26 | 165.78 |
| 26 | 75.62 | 34.99 | 56 | 7.18 | 166.68 |
| 27 | 77.17 | 46.89 | 56 | 4.11 | 167.58 |
| 28 | 78.09 | 61.69 | 57 | 1.04 | 168.47 |

communicate. If the other station is prepared and waiting, you should achieve communications for one minute or longer, even at long distances.

## finding oscar

The AMSAT *Newsletter* publishes a daily listing of a *single* OSCAR 7 LAN, the time for that LAN in GMT, and the orbit number. (The LAN is simply a listing of the geographical location on the equator where OSCAR 7 can be found during a south-to-north pass at one specific time on that day.)

As OSCAR 7 moves in its orbit, the earth rotates under the OSCAR-7 orbit. After 114.9 minutes, OSCAR 7 will again cross the equator. This orbit will cross the equator 28.725 degrees west of the LAN given in the AMSAT listing. On each succeeding orbit of OSCAR 7, a new LAN is obtained, 28.725 degrees west and 114.9 minutes later than the preceding orbit. Thus, the listing given by AMSAT gives you the tool to locate OSCAR 7 on a regular basis.

## the oscar orbit

OSCAR 7 is in a retrograde orbit with an inclination of 101.7 degrees and a period of 114.94 minutes. It makes approximately twelve orbits per day. A retrograde orbit has an inclination greater than 90 degrees as measured from a plane running through the equator. The retrograde orbit also has a very long life. The rate of orbit decay is very low, and the orbit ground track will not change appreciably for several years. Thus, any effort spent in plotting the orbit

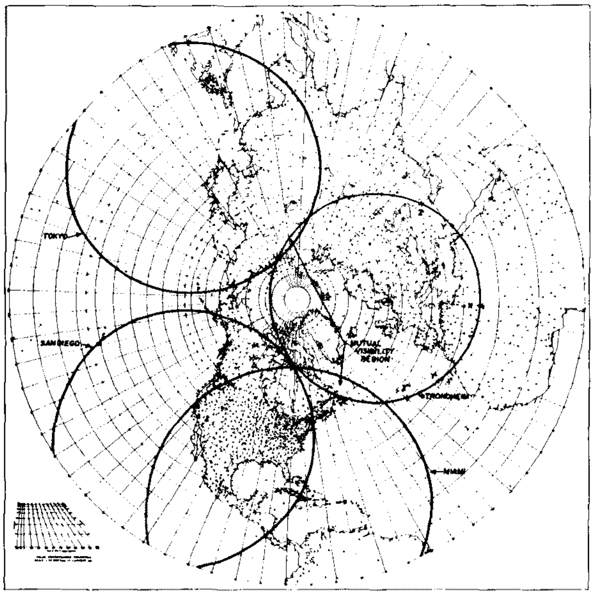


fig. 3. Ground-station mutual visibility regions plotted on a polar-stereographic projection of the northern hemisphere.

will not be wasted. **Tables 1** and **2** give latitude and longitude of OSCAR 7 orbits at one-minute intervals for the northern and southern hemispheres, respectively. This orbit starts at 140 degrees west longitude. (The first half of the orbit was calculated manually.)

The orbit should be plotted onto a clear acetate overlay of the map projection to be used, using drafting ink or other permanent medium. Carefully plot accurate registration marks such as the equator or the north or south pole. Note that the end of the orbit is 168.5 degrees west and 28.5 degrees east of the starting point. This occurs because of the earth's rotation beneath the satellite orbit track. Earth rotation has been accounted for at every instant along the orbit ground track, with the end of the orbit becoming the starting point for the next orbit.

The orbit overlay is used on the polar-stereographic projection with a tack or a pin at the pole, leaving the overlay free to rotate. The Mercator projection overlay must be moved back and forth along the equator. In either case, locate the start of the orbit at the location of the selected LAN. The orbit overlay then locates OSCAR 7 at one-minute intervals for that orbit.

An orthographic-meridional projection of the earth provides a convenient and accurate method for plotting a ground station visibility region. To illustrate the plotting techniques, we'll construct the visibility regions of four ground stations. The first step is to determine the northern most point (NMP) and southern most point (SMP) of the visibility region. To find these regions, you must add to, and subtract from, the *latitude* of each station the earth central half-angle for OSCAR 7, which is 37.5 degrees. It has been assumed that, for each station, there exists a zero-degree horizon and a zero-degree antenna elevation.

The following table shows the NMP and SMP for the four ground stations. Note that the NMP for Trondheim is over the north pole at $63.4 + 37.5 = 200.9$ *degrees* (actually $90 - 10.9 = 79.1$ *degrees*).

| station | latitude (degrees) | longitude (degrees) | NMP (degrees) | SMP (degrees) |
|---|---|---|---|---|
| San Diego | 32.7 N | 117.0 W | 70.2 N | 4.8 S |
| Miami | 25.8 N | 80.2 W | 63.3 N | 11.7 S |
| Trondheim | 63.4 N | 10.4 E | 79.1 N | 25.9 N |
| Tokyo | 35.7 N | 140.0 E | 73.2 N | 1.8 S |

The next step is to plot the NMP and SMP for each station on the orthographic-meridional projection

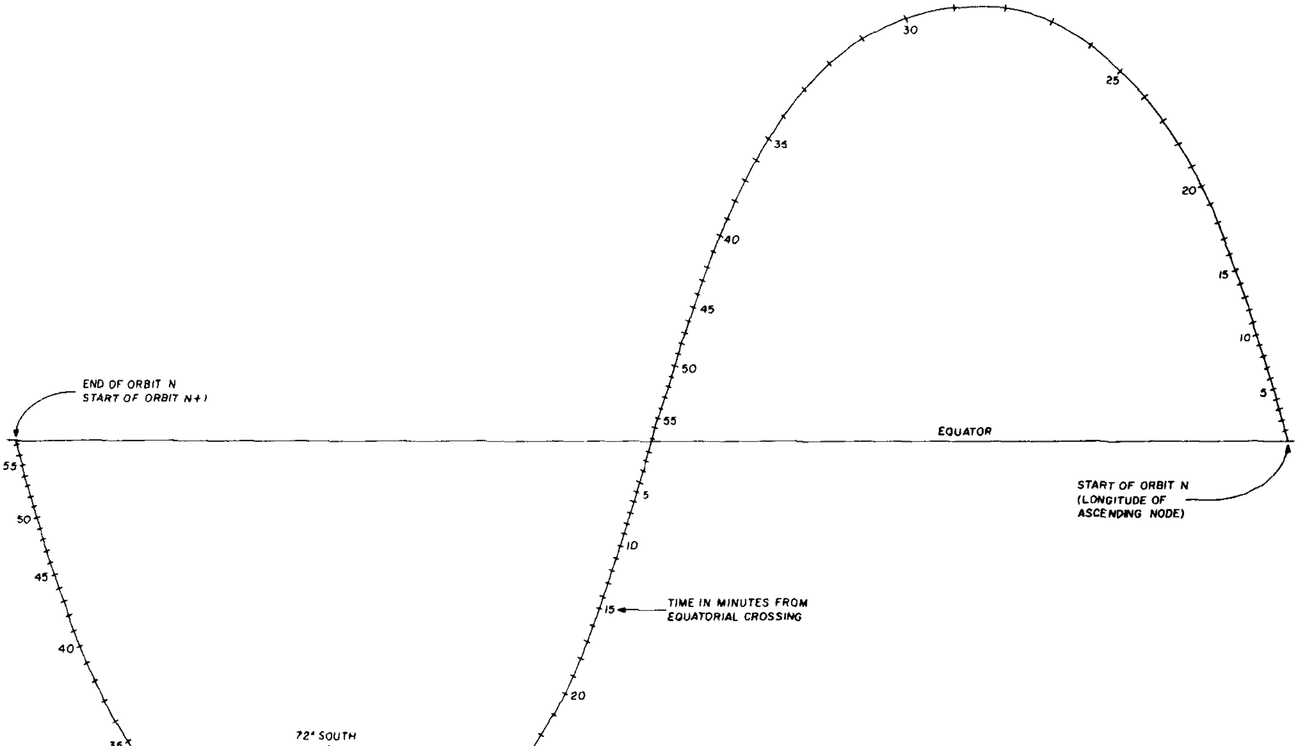
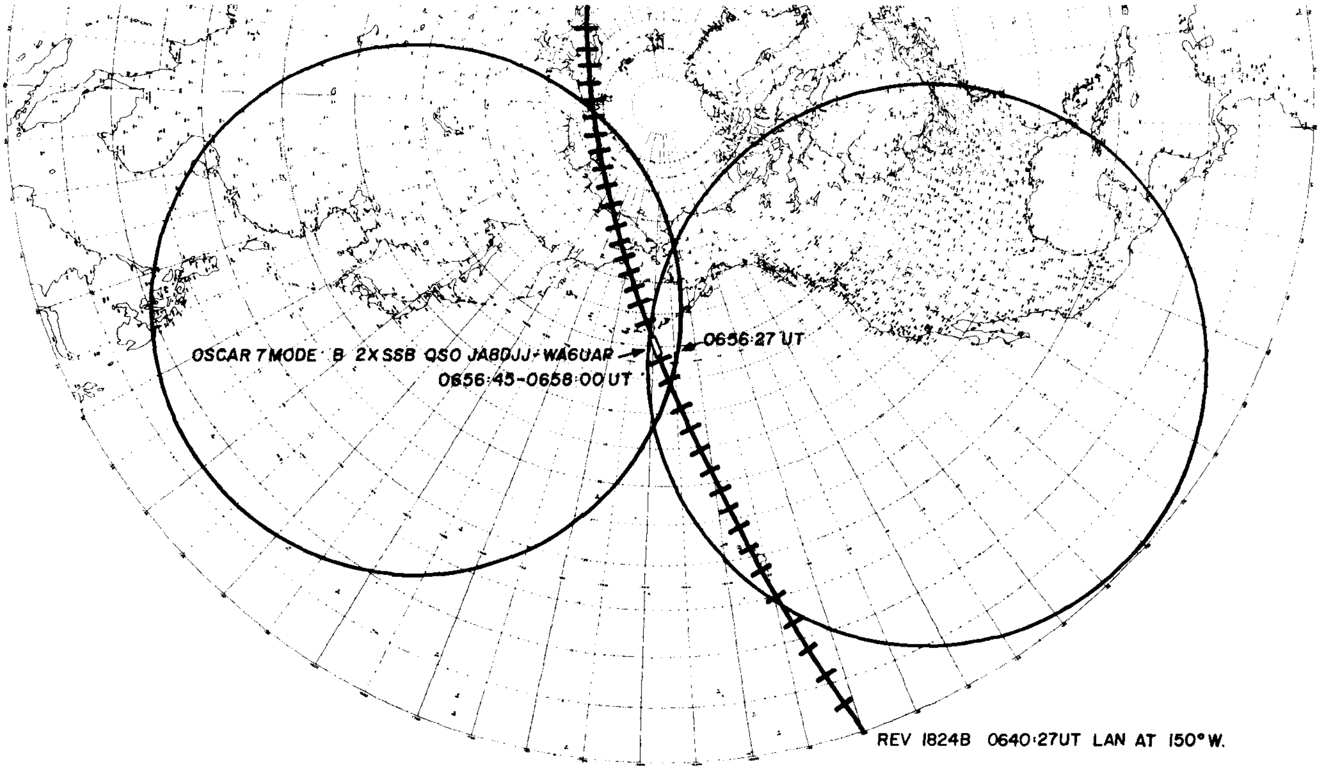



**fig. 4. Single-orbit overlay for a Mercator map projection. The overlay must be slid along the equator until the start of the orbit is on top of the longitude of the OSCAR 7 longitude of ascending node (LAN). The overlay will then show the orbit of the satellite.**




**Plot on polar-stereographic map showing mutual satellite visibility circles of WA6UAP in San Jose, and JA8DJJ in Sapporo, Japan. Two-way ssb communications have been established a number of times between these two stations, but OSCAR 7 is in the proper postion only one or two times per month.**

(**fig. 6**) and connect the points with a straight line. Note that only the NMP and SMP for Trondheim are completed to keep the figure uncluttered for further use. The table of longitudinal corrections is made by recording the value, as read from **fig. 6**, for each increment of latitude. For example, at the SMP (25.9 degrees north) the correction line starts at zero degrees (**A**). At 30 degrees north, the correction line intersects at 26.5 degrees (**B**). Proceeding to 35 degrees north latitude, the line intersects at 40 degrees (**C**). This process is continued until the correction for NMP (180 degrees) is reached. By using smaller increments, a higher degree of accuracy can be obtained for plotting the region of visibility.

To complete the region of visibility, it's necessary to plot the latitude and longitude plus its corrections.

**table 3. Longitude correction values for the four ground-station examples discussed in the text.**

| Trondheim | | San Diego | | Miami | | Tokyo | |
|---|---|---|---|---|---|---|---|
| north latitude (degrees) | correction (degrees) value | north latitude (degrees) | correction (degrees) value | north latitude (degrees) | correction (degrees) value | north latitude (degrees) | correction (degrees) value |
| 79.1 | 180.0 | 70.2 | 0 | 63.3 | 0 | 73.2 | 0 |
| 75.0 | 128.0 | 65.0 | 31.x | 60.0 | 24.x | 70.0 | 25.x |
| 70.0 | 107.0 | 60.0 | 39.0 | 55.0 | 32.0 | 65.0 | 39.0 |
| 65.0 | 94.5 | 55.0 | 43.2 | 50.0 | 37.2 | 60.0 | 45.0 |
| 60.0 | 84.5 | 50.0 | 45.5 | 45.0 | 40.0 | 55.0 | 47.5 |
| 55.0 | 76.0 | 45.0 | 46.1 | 40.0 | 42.0 | 50.0 | 48.5 |
| 50.0 | 67.5 | 40.0 | 46.0 | 35.0 | 42.5 | 45.0 | 48.5 |
| 45.0 | 59.0 | 35.0 | 45.5 | 30.0 | 42.5 | 40.0 | 47.8 |
| 40.0 | 50.0 | 30.0 | 44.0 | 25.0 | 41.8 | 35.0 | 46.5 |
| 35.0 | 40.0 | 25.0 | 42.1 | 20.0 | 40.5 | 30.0 | 44.5 |
| 30.0 | 26.5 | 20.0 | 39.5 | 15.0 | 38.8 | 25.0 | 42.0 |
| 25.9 | 0 | 15.0 | 36.5 | 10.0 | 36.2 | 20.0 | 39.0 |
| | | 10.0 | 32.5 | 5.0 | 32.9 | 15.0 | 35.0 |
| | | 5.0 | 27.0 | 0 | 28.2 | 10.0 | 30.5 |
| | | 0 | 19.0 | 5.05 | 22.3 | 5.0 | 23.5 |
| | | 4.85 | 0 | 10.05 | 12.5 | 0 | 12.5 |
| | | | | 11.75 | 0 | 1.85 | 0 |

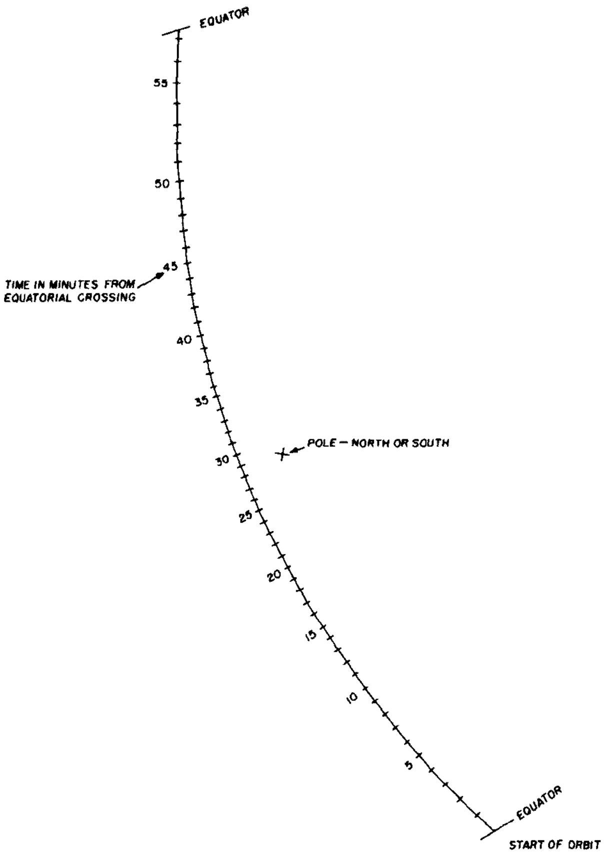



**fig. 5. Single-orbit overlay for a polar-stereographic map projection. Placing the overlay at the longitude of the ascending node (LAN) will show the OSCAR 7 orbital path. The marks indicate the position for each minute after the equatorial crossing.**

Using Trondheim, the longitude is 10.4 degrees east and the latitude of the SMP (0 degrees correction) is 25.9 degrees north (X). Now, referring to **table 3**, the longitude correction for 30 degrees north latitude is 26.5 degrees. This correction is added to, and subtracted from, the initial longitude, *10.4 degrees east + 26.5 degrees = 36.9 degrees east* (**Y**) and *10.4 degrees east − 26.5 = 16.1 degrees west* (**Y′**). Again for 35 degrees north, the longitude points are *10.4 + 40 = 50.4 degrees* (**Z**) and *10.4 − 40 = 29.6 degrees west* (**Z′**). Once the points have been plotted they can be connected with a smooth curve to form the region of visibility.

The procedure described above works for any map projection. For the polar-stereographic projection, however, the visibility region of stations located at or above 37.5 degrees can be plotted using the following shortcut. Locate the NMP and SMP on the longitude line running through the north (or south) pole and your station. Bisect the line connecting the NMP and SMP and use that point as the center of your visibility region. A few seconds with a compass and you'll have your region of visibility. The same technique can be used to plot the 4500 n. mi. (8334 km) range. The only difference is that the earth central half angle is 75.0 instead of 37.5 degrees. This plot should only be done for your own station location to establish the extreme limits of your communications capability.

## antenna elevation

The visibility region discussed previously was based on an antenna elevation of zero degrees (antenna horizontal to the earth's surface). Normally, the amateur using OSCAR 7 and working stations at maximum ranges will be content with this condition. However, if you wish to elevate your antenna, the nomograph of **fig. 7** may be used to assess the effects of antenna elevations up to 15 degrees. If, for example, you elevate your antenna 15 degrees, the earth central half angle will be reduced to 25.25 degrees and your communications range (radius) will be reduced to 1510 n. mi. (2797 km).

## results

Communications with distant stations using antennas fixed in azimuth and elevation has been quite successful. A reproduction of a POPP chart, showing a contact through Mode B between JA8DJJ and WA6UAP, illustrates the usefulness of knowing the mutual visibility area. Knowing the satellite's ground track precisely permits low power experimentation. Numerous contacts have been made through Mode B while operating mobile using a ten-watt output ssb transceiver.

The POPP chart has been in use by the authors only since OSCAR 7 was launched in November, 1974. The basic ideas, methods and applications,

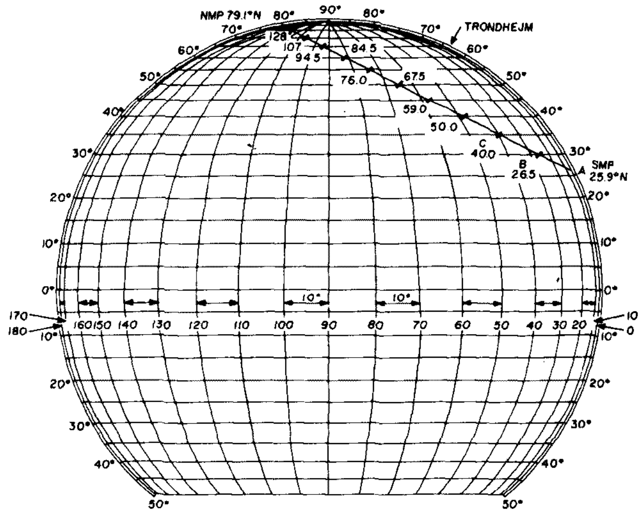



**fig. 6. Orthographic-meridional projection. The longitude corrections are read from the line that connects the northernmost point (NMP) and southernmost point (SMP) of the visibility region.**

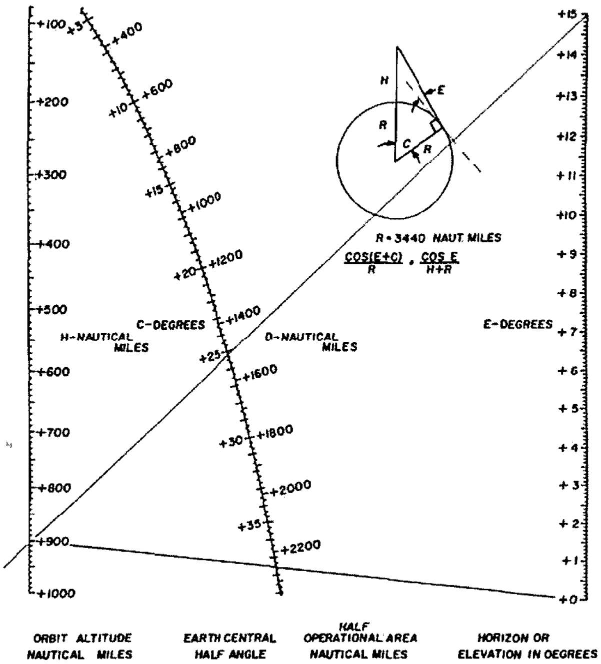



**fig. 7. Nomograph for determining the effects of antenna elevation on satellite communications range. An antenna elevation of zero degrees provides maximum range; about 2250 nautical miles for OSCAR 7 (4500 nautical miles to another ground station). Antenna elevation of 15 degrees gives about 1500 nautical mile half range.**

however, can be applied to any satellite in a circular orbit if the satellite altitude and period are known.

## acknowledgements

We wish to thank Takashi Ishigaki, JA8DJJ, for keeping numerous early morning schedules and Tom Berthold for his assistance with the HP-9100A computer.

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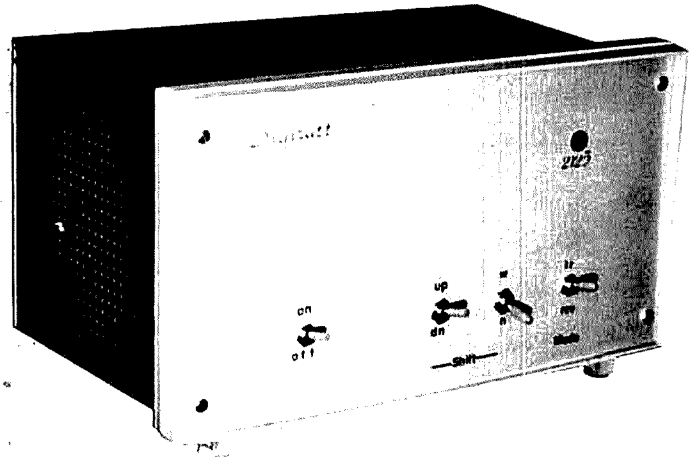

# digiratt —

## RTTY AFSK generator and demodulator

Complete details for a precision AFSK generator and phase locked loop terminal unit

**The *Digiratt* is a complete** terminal unit and AFSK tone generator for use on the vhf bands. The tones are digitally synthesized and will remain accurate to within 0.1 Hertz. The terminal unit is a phase-locked loop unit which is capable of resolving 170-Hz narrow-shift as well as the 850-Hz shift normally found on the vhf amateur bands.

Briefly, the features of the *Digiratt* are,

1. Upward/downward operation by means of front panel switches.

2. Visual indication of 2125-Hz tones by means of an LED. No oscilloscope is needed for proper operation.

3. Built-in test functions for PLL alignment.

4. Constant system performance checking.

And finally, as noted above, long-term tone accuracy to 0.1 Hz in the AFSK generator. Additionally, a PLL terminal unit will inherently follow a drifting signal and copy signals which are missing mark or space information.

### circuit description

The demodulator (U1) compares the incoming frequencies to its internal current-controlled oscillator and generates a digital signal that indicates when the frequencies are identical. Additionally, the IC has provisions to lock the internal oscillator to the incoming signal, provided that signal is within the detection bandwidth. At 2125 Hz the detection bandwidth for this circuit is approximately 220 Hz, or 110 Hz either side of 2125 Hz, which is adequate for 170-Hz shift. At 2975 Hz the detection bandwidth is about plus or minus 135 Hertz. All bandwidth measurements were made with a 300 mV input signal. Although the oscillator can be shifted down below 1400 Hz, the detection bandwidth widens out to plus or minus 198 Hz, and 170-Hz copy becomes doubtful. Readers with this particular application are referred to a 567 specification sheet for bandwidth reduction suggestions.

The output of U1 is low with a detected input, and provisions have been made to invert that signal

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when required. A front panel switch selects either the 567 output or its complement (U2) to drive the selector magnet transistors, thereby resulting in either normal or reverse copy.

The selector magnet drivers, (Q1 and Q2), provide a current return path for the loop supply. This portion of the *Digiratt* is very similar to the NS-1A.[1]

## afsk generator

The precision AFSK generator consists of a master oscillator running at 5.95 MHz. A trimmer is provided for setting the frequency. However, since such a great division of the crystal frequency takes place, the oscillator can be off frequency 100 kHz and the resultant AFSK tone error will be 1 Hz or less. Counter U4 and U5 divide the frequency to 59.5 kilohertz. U6 further lowers the frequency to 5950 Hz and U8 divides it by two, resulting in a symmetrical 2975-Hz square wave at pin 12 of U8. U7 is a programmable counter which divides the 59.5 kHz down to 4250 Hertz. The other half of U8 next divides by two, resulting in a symmetrical 2125-Hz square wave at pin 9 of U8.

Square waves are very rich in harmonics. Therefore, the combination of R11, R12, C8, C9, C10 and C11 are used to turn the square waves into trapezoids. Diodes CR3 and CR4 then smooth the tops, resulting in 1 volt p-p quasi-sinewaves at the output.

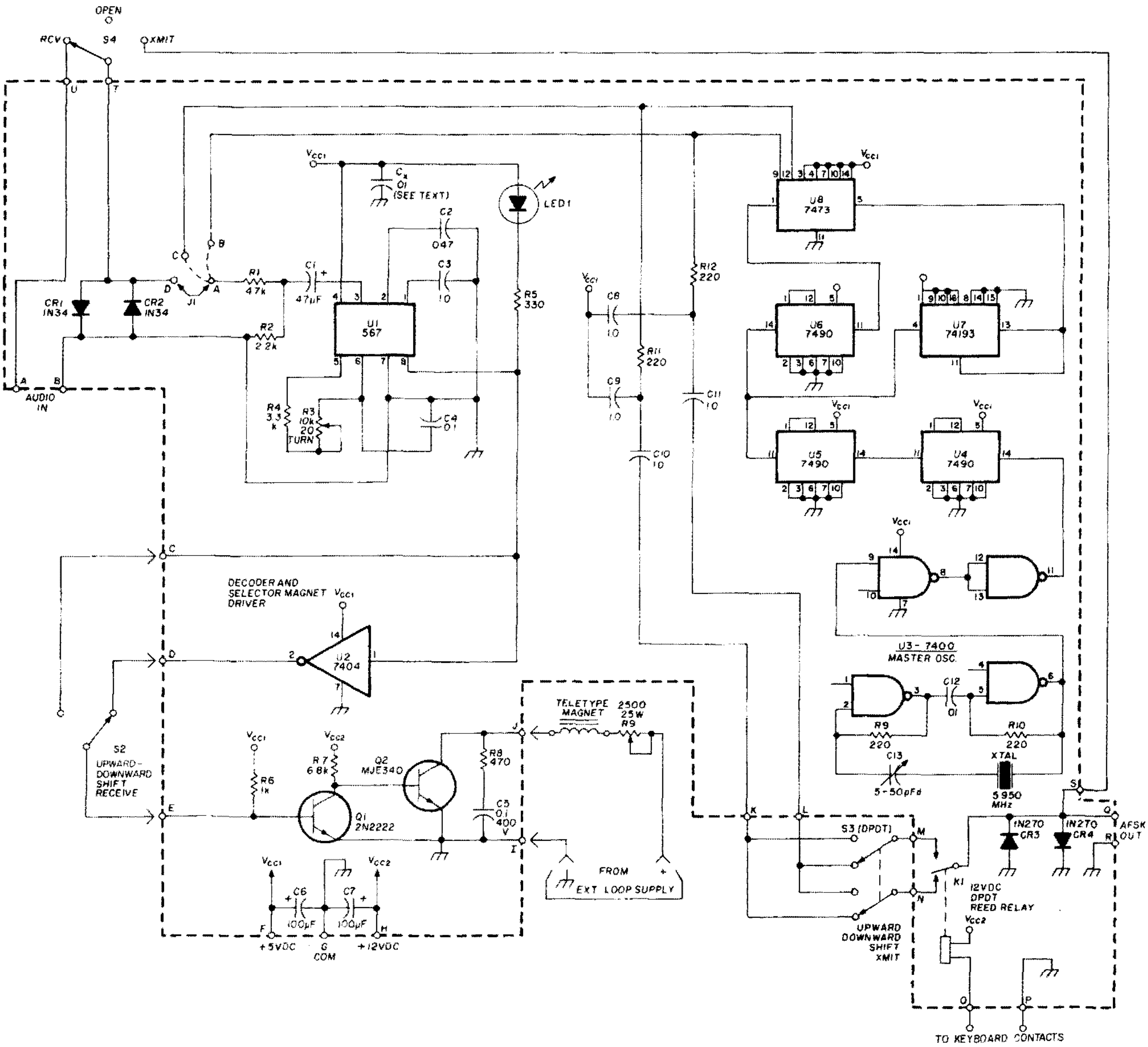

**fig. 1. Schematic diagram of the complete terminal unit. The dashed line indicates the components mounted on a circuit board. K1 is a 12-volt dpdt reed relay. The external loop supply is approximately 100 volts at 100 milliamperes.**

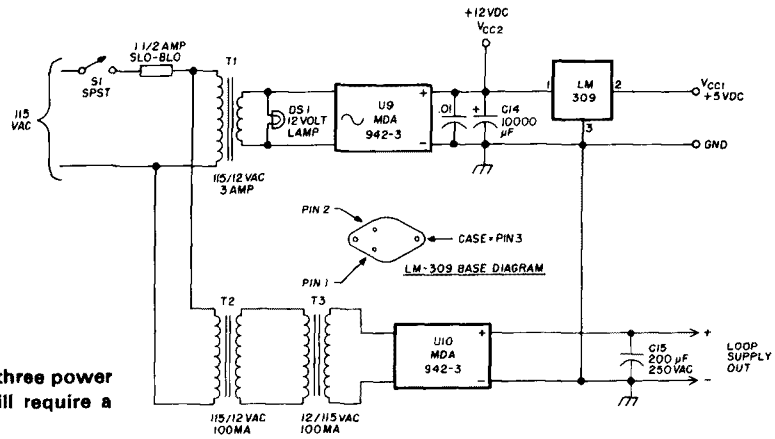


**fig. 2. Schematic diagram of the three power supplies. The 5-volt regulator will require a heatsink.**

If signals above 1 volt p-p are required, a simple transistor amplifier will be sufficient.

The power supplies are straightforward and the only minor precaution is to make sure the LM 309 is mounted on a heatsink. Also, don't fail to install C6 and C7 on the circuit board and bypass U1s pin 4 at the IC with a 0.01 $\mu$F capacitor. You may have to bypass the rest of the ICs at their $V_{CC}$ pins if you run into oscillation problems. A novel feature of the *Digiratt* is LED 1. When jumper J1 is in the A-B position, the PLL oscillator can be adjusted to the same frequency as the precision AFSK generator's 2125-Hz output. This assures correct PLL alignment. When the jumper is in the A-D position, the LED will light when receiving 2125-Hz signals from your receiver. An oscilloscope is not needed because a PLL only cares about the frequency within its bandwidth and disregards the other frequencies all together. This will be explained in more detail in the setup and operation section.

## construction

The prototype is housed in a 5 x 6 x 9 inch (12.7x15.2x22.9cm) cabinet. All power supplies, the loop supply, and the circuit board are mounted within the enclosure. Use care with lead routings; lead lengths should be kept to a minimum and ground loops should be avoided. If you are using the *Digiratt* in a high intensity rf field, use a single point-ground system on the ground returns.

## setup and operation

Set the master oscillator on frequency if you have a frequency counter. If you don't, C13 can be left 50 per cent meshed (the resultant error won't be more than 1 Hz at the output frequencies). Place jumper J1 in the A-B position and turn R3 all the way to one end. When the PLL is locked, LED 1 will come on. The final position for R3 should be midway or be tween the two dropout points.The PLL is now aligned to 2125 Hz and will not require any further adjustments unless 2125 Hz is no longer used. In that case, the PLL will have to be aligned to one of the two new frequencies. For normal amateur use you can set it to 2125 Hz and forget it.

Reconnect the jumper to A-D. Connect the unit to your vhf receiver and tune in an RTTY signal. Flip S2 to the opposite direction if the copy is garbled. That's all there is on vhf where the tones are fixed. For high-frequency use, adjust the receiver until the LED is blinking on and off as the tones shift. Again, flip S2 to clear up the copy. Make sure you have at least 200 mV of audio into the *Digiratt.* Diodes CR1 and CR2 will limit any excessive signals to a level the PLL can handle. Remember to keep the loop supply *off* the keyboard. This system encodes the keyboard output, then detects it to drive the selector magnet. This way you always know the complete system is working even in local loop operation. Also, you shouldn't use the AFSK generator on ssb in the high-frequency bands since the instantaneous switch points between tones contain no modulation. This would result in clicks being transmitted and a possible citation from the FCC.

One final note on the *Digiratt*: if you're a purist at heart and want perfect sine waves at the AFSK output, construct two pi-section filters using 88 mH toroids and place one in each line just ahead of relay K1. Diodes CR3 and CR4 will then have to be removed.

My thanks to Gus, K9FUI, who shared his considerable knowledge of *Teletype* machines with me and also to my wife, Donna, for her patience while I worked many late nights on the project.

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**ham radio**

# pi network design and analysis

A new approach to the design of pi networks for amateur transmitting equipment which allows the designer to determine how the network will perform at different frequencies, with varying antenna loads

**Simplified procedures** for the design of pi networks have appeared in numerous magazine articles and are included in the *ARRL Handbook* and other reference books. No equation solving is required with these procedures since the design data is provided in terms of inductance and capacitance which the user reads directly from tables or curves.[1-6] All of these data are based upon the three classic reactance equations for the design of pi networks.

However, the simplicity afforded by these procedures is not obtained without penalty. Only design center information is provided, and in some cases the network may not perform as expected. Characteristics which are of prime importance to the amateur, *e.g.,* the usable tuning range, and the tolerable limits of load variation due to the vswr of the antenna system, cannot be evaluated so the serious designer is handicapped. It is not unusual to encounter circumstances where the loading control exhibits less than normal effectiveness, and it is difficult or impossible to obtain normal plate current loading. Some contemporary commercially designed equipment, as well as homebrewed equipment, will exhibit this condition unless the vswr of the transmitter load is kept very close to 1:1.

Prior amateur literature is virtually void of discussion on the behavior of pi networks when the operating conditions depart from the values used for the initial design of the network. More often than not such departure is normal in amateur operations; therefore, an investigation along these lines appeared to be a project worth pursuing. Some unexpected and very interesting results came from this work, including the development of a new procedure for the design and performance evaluation of pi networks, and discrete criteria for determining when supplemental impedance transformation must be used.

These developments were based upon two relatively obscure equations which also provided the *seed* from which other equations used in the procedure were derived. Compared with contemporary design methods, more work is involved in the application of the new procedure, but the benefits of greatly increased design control, and the addition of a means for performing detailed performance analysis, are substantial compensation for the added effort.

The objectives of this article are twofold: first, to describe the practical application of the new design procedure, and secondly, to provide some insight into the analysis which led to its development. Readers who shy away from the mathematical discussions should note that numerical solutions of all the equations in this article require only basic arithmetic and square-root operations — the modern, hand-held calculator is completely adequate for designing a new pi network, or for analyzing an existing one.

It should be understood at the outset that no lack of generality is inferred by having oriented the following discussions in terms of networks for use with vacuum-tube amplifiers. This approach was chosen because of the widespread application of tube amplifiers, and the probable continued use of vacuum tubes in the final stages of amateur transmitters for some time to come. However, readers should find it amply apparent that the equations and design procedures presented in this article can be applied to the design of any lowpass pi network where there is no provision for adjustment of the inductive branch.

## design equations

The basic pi network is shown in **fig. 1A**. When used for coupling a low impedance load to a vacuum

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tube, $C1$ is the loading control and $C2$ is the plate tuning control, $R1$ represents the external load impedance, and $R2$ is the required plate load resistance as measured at terminals 1 and 2. In network design and analysis work, however, reactance terms must be used in place of inductance and capacitance, and the component designators are modified as shown in **fig. 1B**. With very few exceptions *reactance* terms are used throughout the text of this article.

**Equations 1, 2** and **3**, or minor variations thereof, are the three classic pi network design equations upon which the simplified design procedures are based

$$X_{C2} = \frac{R2}{Q} \tag{1}$$

$$X_{C1} = R1\sqrt{\frac{R2}{R1(Q^2+1) - R2}} \tag{2}$$

$$X_L = \frac{QR2 + \dfrac{R1R2}{X_{C1}}}{Q^2+1} \tag{3}$$

These equations are used to calculate the three network reactances when appropriate values are substituted for $R1$, $R2$, and $Q$. Normally the nominal antenna load impedance of 52 ohms is substituted for $R1$; $R2$ is the required plate load resistance; and $Q$ is selected in the range from 10 to 20.

## analysis considerations

In class-AB1 and class-B linear amplifiers the dc plate current, and the linearity of the amplifier, depend upon the value of plate load resistance. Thus the optimum value of plate load resistance is coincident with a specific value of plate current to which the amplifier must always be adjusted; this load must also be non-reactive. In a properly neutralized amplifier the latter requirement is obtained by tuning to the center (minimum current point) of the plate current dip. The adjustment procedure consists of alternately adjusting $C1$ and $C2$ until the specified value of plate current is obtained at the dip point.

Since capacitors $C1$ and $C2$ are the only network adjustments, they must permit the optimum value of plate load resistance to be maintained regardless of changes in operating frequency and/or changes in antenna load impedance. Thus, an analysis of network performance with respect to any external influence which will effect its adjustment must be made in terms of the requirements imposed upon $C1$ and $C2$.

It might be expected that **equations 1** and **2** could be used for this purpose. However, their use is precluded by the presence of $Q$ in both equations, and the lack of $R1$ as a factor in **equation 1**. $Q$ is not an independent variable and therefore does not remain constant as the frequency and/or the value of $R1$ is changed. The following equations are elegantly suited to the desired analysis because they are totally free of such limitations

$$X_{C1} = \frac{R1X_L}{R1 + \sqrt{R1R2 - X_L^2}} \tag{4}$$

$$X_{C2} = \frac{R2X_L}{R2 + \sqrt{R1R2 - X_L^2}} \tag{5}$$

In these equations the reactance of $C1$ and $C2$ is stated in terms of the independent variables $R1$ and $X_L$; $R2$ is treated as a constant. These equations were first derived by classical methods, and can be found in the **1932** edition of Everitt's *Communications Engineering*.[7] It is strange that these equations have been available for more than 40 years but have been neglected in the amateur literature except for a footnote reference by George Grammer in *QST* in 1957.[3]

Since the pi network is usually used to provide a

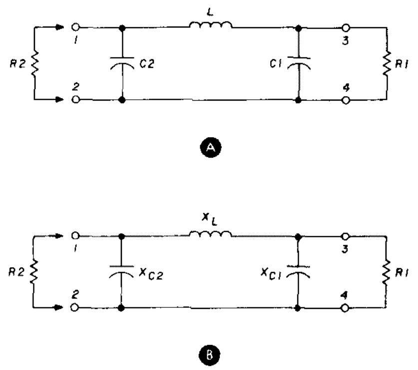


**fig. 1. Basic pi network used for impedance matching in rf power amplifiers. In *A* $R1$ is the load impedance (usually 52 ohms), $R2$ is the plate load impedance, $C1$ is the loading control, $C2$ is the tuning control, and $L$ is the fixed inductor. In the circuit in *B* the inductor and capacitors are shown as reactance terms.**

match between two dissimilar impedances, its basic symmetry is easily overlooked, and the identical forms of **equations 4** and **5** may be surprising. They are markedly different from **equations 1** and **2**, just as **equations 1** and **2** have no resemblance to each other and give no clue to as to the symmetry of the pi network.

Curves showing the relationships expressed in **equations 4** and **5** are plotted in **fig. 2**. These curves are based upon an arbitrary prototype network designed to provide a plate load resistance of 2500 ohms and to accommodate variations in output load impedance in a nominal 52-ohm antenna system. The data for plotting the curves was obtained from repeated solutions of **equations 4** and **5** with a value of 200 ohms for $X_L$, and $R1$ varied incrementally

from 16 ohms (the value of $R1A$ in this case) to over 100 ohms.

Frequent reference to these curves, and particularly the points designated as $R1A$ and $R1B$ , will occur throughout the following text. Equations for the determination of the value of $R1$ at these points, and a discussion of the significance of each point will be covered later.

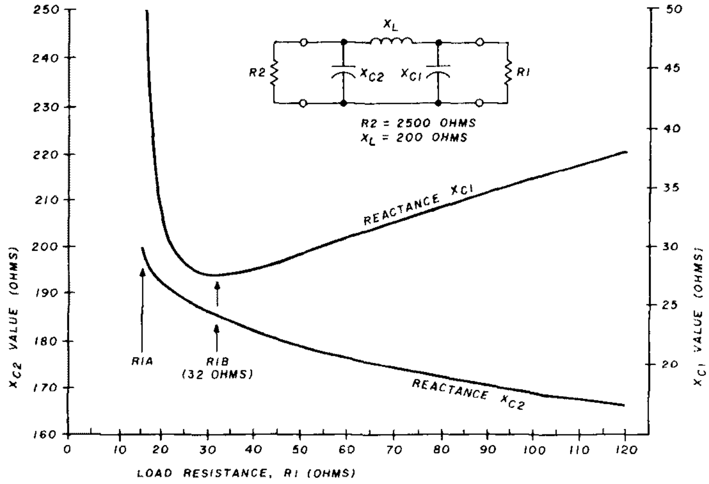


**fig. 2. Plot of capacitive reactance $X_{C1}$ and $X_{C2}$ vs load resistance, $R1$. Note that the $X_{C1}$ curve is double valued; for most designs values below point $R1B$ should be avoided (see text). Slope of reactance curve above $R1A$ is excessive and may result in critical tuning.**

Comment on the curve of $X_{C2}$ (plate tuning) is necessarily brief since it has no unusual or critical characteristics. Operation is smooth and continuous throughout its required tuning range, and the moderate change in tuning rate is of no consequence. No operational problems should exist as long as sufficient capacitance range is available.

Cause for design and operational problems with the pi network is apparent in the characteristics revealed by the curve $X_{C1}$. In terms of $R1$, the $X_{C1}$ curve is double valued for all values of $X_{C1}$ except the minimum. These dual values of $R1$ cause two mutually inverted calibrations to exist for the loading control, effectively dividing the $R1$ range into two sections. Division occurs at the point $R1B$ so that all values of $R1$ less than $R1B$ may be defined as being in the *minor section*, while all values of $R1$ above and including $R1B$ may be defined as being in the *major section*. It is preferable to design the network so operation is confined to only the *major section*. When this is done, ambiguity is avoided with respect to the direction of loading control rotation to obtain increased loading, there is no confusing area of dual and inverted calibrations, and the tuning rate of the loading control does not change drastically.

The minimum point of the $X_{C1}$ curve locates the value of $R1$ below which undesirable changes in the network's operating characteristics begin to take place. If a sufficiently low value of $R1$ cannot be obtained at $R1B$, the designer must either use a supplemental impedance transformation, or resort to a compromise design with a part of the *minor section* included in the operating range. The latter option is not an unconditional alternative since only limited extension of the $R1$ range can be obtained in this way. The conditions associated with this limitation, and a design criteria for applying the compromise design, will be discussed later.

## significance of $R1A$ and $R1B$

Mathematical proof can be given, but it will suffice to state here that solutions of **equations 4** and **5** are valid for this analysis only when the quantity under the radical sign is equal to or greater than zero. Since the plate load resistance, $R2$, must remain constant and the $R1R2$ product cannot be less than $X_L{}^2$, $R1$ is limited accordingly to a critical minimum value. An equation defining the critical value of $R1$ is obtained when the expression under the radical sign is equated to zero, and then solved for $R1$.

$$R1R2 - X_L{}^2 = 0 \tag{6A}$$

$$R1R2 = X_L{}^2 \tag{6B}$$

$$R1\ (\text{critical}) = R1A = \frac{X_L{}^2}{R2} \tag{6C}$$

From these equations it can be seen that whenever $R1$ is less than $R1A$ the network cannot be adjusted to provide the required value of $R2$; the maximum plate-current loading, therefore, will be less than normal.

It is worth noting that a unique but seldom used condition exists when the product $R1R2$ is equal to $X_L{}^2$, causing the radical term to become zero. The remaining $R1$ and $R2$ terms cancel out in **equations 4** and **5**, with the result that both $X_{C1}$ and $X_{C2}$ equal $X_L$! This is the familiar situation when the pi network behaves like a one-quarter wavelength section of transmission line with a characteristic impedance

equal to $X_L$. This form of the pi network has been used in cascaded pairs to form an effective narrowband harmonic suppression filter.

The value of $R1$ at the minimum point of the $X_{C1}$ curve is designated $R1B$, and as the following derivation will show, it is always twice the value of $R1A$. Subsequent discussions will also show that $R1B$ is a major factor influencing the design of any pi network. Since, by definition, $R1B$ is the value of $R1$ at the minimum point (point of zero slope of the $X_{C1}$ curve), the derivation of the equation for $R1B$ involves the application of differential calculus. Thus the prerequisite slope equation* (**equation 7**) is derived from **equation 4**, then equated to zero and solved for $R1$ to obtain the equation for $R1B$. Initially **equation 7** appears somewhat formidable, but its terms contain only three factors, and considerable cancellation may be performed when it is equated to zero and solved for $R1$.

$$S = \frac{X_L(R1R2 - 2X_L^2)}{[2\sqrt{R_1R_2 - X_L^2}\,][R_1 + \sqrt{R_1R_2 - X_L^2}\,]^2} \qquad \textbf{(7)}$$

where $S$ = slope of the $X_{C1}$ *curve. Substituting zero for S*, and solving for $R1$†,

$R1$ (at minimum point of $X_{C1}$ curve) =

$$R1B = \sqrt{\frac{2X_L^2}{R_2}} \qquad \textbf{(8)}$$

## factors $T_r$ and $K$

Both $T_r$ and $K$ are simple dimensionless ratios which are inversely proportional to the value of $R1B$; since they are important factors in some of the following equations, it's important to define them.

**1.** $T_r$ is the ratio of $R2$ to $R1B$ ($R2$ divided by $R1B$); therefore $T_r$ represents the transformation ratio of the network when, and only when, $R1 = R1B$.

**2.** $K$ is the ratio of $R1$ to $R1B$ ($R1$ divided by $R1B$); it is the value of R1 normalized with respect to $R1B$. Values of $K$ below 1.0 are in the *minor section* of the network, whereas values of $K$ of 1.0 or above are in the *major-section* of the network. $K$ can never be less than 0.50, and although there is no maximum limit for $K$, it will seldom be greater than 6.0.

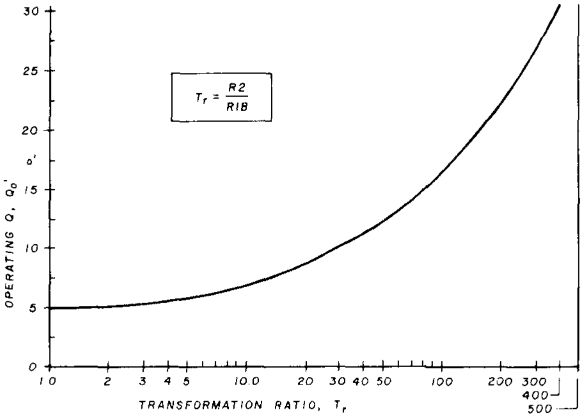


**fig. 3. Operating $Q$ ($Q_o$) of a pi network versus the impedance transformation ratio, $T_r$.**

The dominant component of the pi network is the series-connected reactance. In transmitter output circuits this component is usually an inductor so the network functions as a lowpass filter as well as an impedance transformer, and useful attenuation of harmonic energy is obtained. An equation for determining the required reactance of this inductor may be derived by keeping the definition of $T_r$ in mind and re-arranging **eq. 8**.

$$X_L = \frac{R2}{\sqrt{2T_r}} \qquad \textbf{(9)}$$

Inasmuch as the value of $R2$ (plate load resistance) is fixed by the design specifications, $T_r$ is the only independent variable in this equation which the designer can use to control the value of $X_L$. However, this control by way of selecting the value of $R1B$, and $R1B$ must be limited to only those values which will place the operating $Q$ of the network ($Q_o$) within the range of 10 to 20. These limits represent a prudent compromise between harmonic attenuation and network efficiency. Harmonic attenuation deteriorates rapidly as $Q_o$ falls below 10; network losses become excessive as $Q_o$ approaches 20. Roughly speaking, changes in $Q_o$ between these limits is accompanied by an 8 dB change in harmonic attenuation, and a 3 dB change in power loss. For the general case the operating $Q$ of a pi network is defined as

$$Operating\ Q = Q_o = \frac{R1 + R2 + 2\sqrt{R1R2 - X_L^2}}{X_L} \qquad \textbf{(10)}$$

*Derivation of this and other equations is available from *ham radio* upon receipt of a self-addressed, stamped envelope.

†Theoretically, in any given pi network $R1$ can be *increased* without limit. In practice, however, excessive circuit losses ultimately restrict the maximum usable value. As $R1$ is increased, $Q_o$ also increases, causing the ratio of unloaded to loaded $Q$ to become progressively lower, and more power to be dissipated in the inductor. The effect is shown in the following data which is based upon the prototype network, and the assumption of an unloaded $Q$ of 300. As a general rule it is advisable to avoid values of $R1$ greater than six times $R1B$.

| R1 | $Q_o$ | power lost | |
|---|---|---|---|
| 52 | 15.76 | −0.23dB | 5.25% |
| 100 | 17.58 | −0.26dB | 5.86% |
| 200 | 20.28 | −0.30dB | 6.76% |
| 500 | 26.00 | −0.39dB | 8.67% |
| 1000 | 33.18 | −0.51dB | 11.06% |
| 5000 | 72.80 | −1.21dB | 24.27% |

This equation can be used to compute $Q_o$ for any value of $R1$ within the operating range of pi network.

Note that **equation 10** differs considerably from the equation, $Q = R2/X_{C2}$, shown in the handbooks.* Errors as great as minus 50% will result when the handbook equation is used to compute $Q$ for pi networks having a very low impedance transformation ratio; errors between 6% and 20% can be expected for pi networks commonly used in vacuum-tube *linear* amplifiers. An example of typical errors may be seen in the following data calculated for the prototype network at three values of output load resistance.

| | | | |
|---|---|---|---|
| Load resistance, $R1$ | **26** | **52** | **104** |
| $Q_o$ (from **eq. 10**) | 14.21 | 15.76 | 17.71 |
| $Q$ (**eqs. 1, 2, 3**) | 13.29 | 14.00 | 14.85 |
| Error | − 6.48% | − 11.17% | − 16.18% |

For the particular case where $R1$ is equal to $R1B$, the operating $Q$ is designated as $Q_o'$

$$\text{Operating } Q \text{ at } R1B = Q_o' = 2 + \sqrt{2(T_r + \frac{1}{T_r} + 2)} \qquad \textbf{(11A)}$$

where $T_r = R2/R1B$. When $T_r$ is equal to or greater than 5, the following equation can be used with less than 1% error.

$$Q_o' = 2 + 2\sqrt{\frac{T_r}{2} + 1} \qquad \textbf{(11B)}$$

**Equation 11** is unique, and indeed remarkable, because it shows that $Q_o'$ is dependent *only* upon the transformation ratio, $T_r$, a value readily obtained by dividing $R2$ by $R1B$. Therefore, a graph of **equation 11** is universally applicable to all pi networks that fall within the range of the scales chosen for $Q_o'$ and $T_r$! Thus the graph of **fig. 3**, or **equation 11**, is invaluable for determining a value for $T_r$ that will result in an acceptable value of $Q_o'$. This value of $T_r$ is then used in **equation 9** to obtain the value for $X_L$.

However, if it is found that $R1B$ must be greater than desired, and the difference is too great to consider the application of a compromise design, it indicates that the pi network by itself cannot satisfy the design requirements, and supplemental impedance transformation must be used. Values of $T_r$ high enough to cause this condition are likely to occur in the design of pi networks intended for use with tubes that require a high plate load resistance such as the 4-1000A. In these cases a pi network by itself will usually fail to provide matching capability for loads much below 52 ohms, since the 52-ohm load point may already be well within the network's *minor section.*

*The error due to the design procedure using **equations 1, 2,** and **3**. These equations treat the pi network as two back-to-back L-networks. The selected value of $Q$ applies only to the input section. It can be shown that the overall operating or loaded $Q$ of the network is the sum of the input section $Q(R2/X_{C2})$ and the output section $Q(R1/X_{C1})$. Therefore, the operating $Q$ will always be higher than the selected $Q$ when using this design procedure.

**Editor**

## pi network design

A step-by-step procedure for the design of pi networks, based on the technique discussed in the previous text, will clarify any questions you may have about this method. A pi network will be designed to match a plate load resistance of 2500 ohms to a nominal 52-ohm system, and will accommodate antenna load variations which might be expected in an amateur station. Operating $Q$ is not specified since it is standard practice to keep $Q_o$ within the range from 10 to 20, and to favor the lower values of $Q_o$ in the interest of circuit efficiency.

**Preliminary considerations**. When a maximum vswr of 2:1 is expected, and the nominal impedance of the antenna system is 52 ohms, the load on the pi network can be any value from 26 to 104 ohms. This impedance range can be covered in the *major section* of a network if $R1B$ is chosen as 26 ohms, and $Q_o'$ does not exceed 15. This limitation on $Q_o'$ is required if $Q_o$ is not to exceed the limit of 20 at the maximum value of R1 (104 ohms). In the following procedure the **steps 1** and **2** will either confirm that $R1B$ can be 26 ohms or provide a value which will enable the $Q_o$ limitation to be maintained.

**1.** Determine value of $Q_o'$ for desired value of $R1B$; use **fig. 3** or **eq. 11**. (For prototype network $T_r = 96.15$ when $R1B$ is 26 ohms, thus from **fig. 3**, $Q_o' = 16$).

**2.** If $Q_o'$ from **step 1** is greater than 15, the value of $R1B$ is too small. Increase the value of $R1B$ by an appropriate increment and repeat **step 1**; repeat **steps 1** and **2** until $Q_o'$ is less than 15. For the prototype network the value of $R1B$ was increased in 2-ohm steps, yielding the following values for $Q_o'$:

| $R1B$ | $T_r$ | $Q_o'$ |
|---|---|---|
| 26 | 96.15 | 16.00 |
| 28 | 89.29 | 15.51 |
| 30 | 83.33 | 15.06 |
| 32 | 78.125 | 14.66 |

**3.** Compute the value of $K$ for the minimum value of $R1$. If the value of $K$ is less than 0.8, see the discussion under *compromise design* (page 36). For the prototype design, $K = 0.81$.

**4.** Compute the value of $X_L$. Use $T_r$ from **step 2** in **eq. 9**. For the prototype network $X_L = 200$ ohms since $T_r = 78.125$.

**5.** Use **eq. 10** to compute $Q_o$ for the upper and lower

**table 1. Twelve complex impedance values which fall on a 2:1 vswr circle plotted on a Smith chart (fig. 4), and the required pi-network reactance values required to provide a match to a plate load resistance of 2500 ohms.**

| point on 2:1 vswr circle (fig.4) | worst-case antenna load impedance | | equivalent parallel components | | required $X_{C1}$ | required $X_C$ of loading control, C1 | required $X_{C2}$ |
|---|---|---|---|---|---|---|---|
| | A | B | C | D | E | F | G |
| 1. | 26.00 | j 0.0 | 26.00 | + inf. | 28.24 | 28.24 | 188.10 |
| 2. | 27.38 | −j 10.27 | 31.23 | −j 83.27 | 27.59 | 41.27 | 185.52 |
| 3. | 32.00 | −j 20.78 | 45.50 | −j 70.05 | 28.70 | 48.62 | 180.40 |
| 4. | 41.60 | −j 31.20 | 65.00 | −j 86.67 | 31.33 | 49.06 | 175.44 |
| 5. | 59.43 | −j 38.60 | 84.50 | −j 130.10 | 33.91 | 45.87 | 171.60 |
| 6. | 86.60 | −j 32.47 | 98.77 | −j 263.40 | 35.68 | 41.27 | 169.21 |
| 7. | 104.00 | −j 0.0 | 104.00 | + inf. | 36.30 | 36.30 | 168.40 |
| 8. | 86.60 | +j 32.47 | 98.77 | +j 263.40 | 35.68 | 31.42 | 169.21 |
| 9. | 59.43 | +j 38.60 | 84.50 | +j 130.10 | 33.91 | 26.90 | 171.60 |
| 10. | 41.60 | +j 31.20 | 65.00 | +j 86.67 | 31.33 | 23.01 | 175.44 |
| 11. | 32.00 | +j 27.80 | 45.50 | +j 70.05 | 28.70 | 20.36 | 180.40 |
| 12. | 27.38 | +j 10.27 | 31.23 | +j 83.27 | 27.59 | 20.73 | 185.52 |

A. Resistive component of the load impedance.
B. Reactive component of the load impedance.
C. Equivalent parallel resistance of the load impedance.
D. Equivalent parallel reactance of the load impedance.
E. Required reactance of $X_{C1}$ (based upon resistance value in column C).
F. Required reactance of the loading control capacitor (obtained by combining values shown in columns D and E).
G. Required reactance of $X_{C2}$ (based upon resistance value in column C).

limits of the $R1$ range. If the values of $Q_o$ are not within the range from 10 to 20, use one of the following alternatives:

**A.** Determine whether or not a modification of the value of $R1B$ will resolve the problem.

**B.** Use the design as is but be aware of its limitations and change operating specifications accordingly.

**C.** Use the design as is and use an antenna tuner to limit the vswr at the transmitter.

**D.** Redesign the pi network for use with supplemental impedance transformation.

For the prototype network $Q_o = 14.21$ when $R1$ is 26 ohms, and $Q_o = 17.71$ when $R1$ is 104 ohms.

**6.** Compute the values of $X_{C1}$ and $X_{C2}$ for the upper and lower limits of the $R1$ range; use **eqs. 4** and **5**. For the prototype network see **table 1**, lines 1 and 7, columns E and F.

**7.** Compute the capacitance values at the desired frequency of operation using the values of $X_{C1}$ and $X_{C2}$ from **step 6**. See **table 1** for prototype network values.

**8.** Compute the inductance value for the desired operating frequency using the value of $X_L$ found in **step 4**. For the prototype network $L = 8.51 \mu H$ when $f_o = 3.74$ MHz, the geometric center frequency of the amateur 80-meter band.*

Pi networks that operate at low values of $Q_o$ are frequently used as input and output circuits for *solid-state* devices. These are fixed-tuned networks designed for singular values of $R1$ and $R2$, and a particular value of $Q_o$ to exist at $f_o$, the geometric center frequency of the network. The design procedure for this type of network is appreciably less complex than that for tunable networks, and **equation 12** below is particularly useful since $X_L$ can be determined directly as a function of $R1$, $R2$, and $Q_o$. The values of $X_{C1}$ and $X_{C2}$ then follow directly from the use of **equations 4** and **5**. The design should not be considered complete, however, until the value of $K$ for $R1$ has been computed and judged to be acceptable. The value of $R1B$ which is needed for this calculation is found with **equation 8**.

Values of $K$ that are not less than 0.70 are acceptable for fixed-tuned networks, whereas values of $K$ that are less than 0.65 should definitely be avoided; otherwise, the network will be difficult to adjust. When the value of $K$ must be increased, it can only be done by an increase in the value of $Q_o$.

$$X_L = (R1 + R2)\ \frac{Q_o^2 + \sqrt{Q_o^2 - \left(Q_o^2 + 4\right)\left(\frac{R2 - R1}{R1 + R2}\right)^2}}{Q_o^2 + 4} \quad \textbf{(12)}$$

## effects of antenna load impedance

Load impedance matching capability is succinctly specified for transmitters in terms of a maximum allowable vswr and a system characteristic impedance. In most cases the values specified for amateur equip-

*A series of HP-25 programs for designing pi-networks, plotting frequency responses, checking the impedance matching range, and selecting inductance and capacitance values are available for $2.50 postpaid from Calculator Design, Box 429, Hollis, New Hampshire 03049.

ment are a vswr of 2:1 and a system impedance of 52 ohms. Impedance values from 26 to 104 are therefore permissible at the transmitter's output terminals, and the impedance will usually be complex (containing both reactance and resistance). Specific values are readily determined with a Smith chart on which a 2:1 vswr circle has been inscribed; the output network of the transmitter should be capable of transforming any impedance whose coordinates are located on or within this circle to a pure resistance equal to $R2$.

Matching is most difficult for the loads associated with the highest vswr values; therefore, the loads with coordinates located on the circle constitute the worst-case requirements of the design specification. There are an infinite number of impedance coordinates located on the 2:1 vswr circle, so it's impractical to evaluate network performance for all of them. However, an adequate evaluation can be performed by using the twelve values of impedance whose coordinates are located by the dots on the vswr circle shown in **fig. 4**. The resistive and reactive components of these twelve impedance values and their equivalent parallel components, are included in **table 1**.

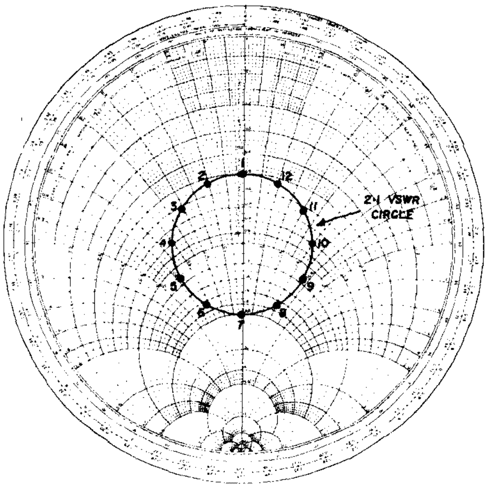


**fig. 4. Normalized Smith chart plot of worst-case load impedance for a 2:1 vswr. These impedance points are used in the evaluation of the pi network (see table 1).**

The equivalent parallel components are included because they are essential for computing the effect of each load impedance on the adjustment of the pi network. And since these components are in parallel with the loading control capacitor, $C1$, the effective parallel resistance is the value of $R1$ loading the network, and the effective parallel reactance must be combined with the reactance of the loading control capacitor so the net value is equal to the required value of $X_{C1}$. Using the prototype network as an example, calculations were made to determine the required value of $X_{C1}$, $X_{C2}$, and the reactance of the loading control capacitor for each of the worst-case load impedances; these data complete **table 1**. The *required capacitance ranges of the plate-tuning and* loading-control capacitors can be computed from the minimum and maximum values of reactance contained in columns F and G of **table 1**. Based upon the geometric center frequency of the 75-meter band (approximately 3.74 MHz) these ranges are

| | |
|---|---|
| Plate tuning | 226 to 253 pF |
| Loading Control | 867 to 2089 pF |

Additional calculations can be made to prepare charts similar to **table 1** for the low- and high-frequency band edges. In these calculations the value of $X_L$ must be changed in proportion to the change in frequency; thus $X_L$ becomes 186.92 ohms at 3.5 MHz, and 214.0 ohms at 4.0 MHz. Using these values of $X_L$ in **equations 4** and **5**, the tabulated reactance data results in the following final capacitance ranges

| | |
|---|---|
| Plate tuning | 196 to 290 pF |
| Loading Control | 705 to 2395 pF |

Since the operating $Q$ changes as the frequency and/or load resistance changes, it is advisable to check the value of $Q_o$ at different frequencies and load values. Using **equation 10**, the values computed for the prototype design are

| frequency | 26-ohm load | 104-ohm load |
|---|---|---|
| 3.50 MHz | $Q_o = 15.37$ | $Q_o = 19.01$ |
| 3.74 MHz | $Q_o = 14.21$ | $Q_o = 17.71$ |
| 4.00 MHz | $Q_o = 13.10$ | $Q_o = 16.49$ |

## compromise designs

A compromise design of the pi network was defined previously as extending the operating range into the *minor-section*. Although this mode of operation is less than ideal in some respects, its objectionable characteristics can be minimized by proper design; in addition, it may offer the only means for meeting some specifications that would otherwise require the use of supplemental impedance transformation.

To understand the limitations that restrict the extent to which the *minor-section* can be used, refer to **fig. 2** and the curve of $X_{C1}$. The slope of the $X_{C1}$ curve progressively increases in the *minor-section* so that virtually all tuning control is ultimately lost as $R1$ approaches the value of $R1A$. Although it is technically valid for $R1$ to have any value down to and including $R1A$, the minimum value of $R1$ must be limited if the region of excessive slope is to be avoided.

Evaluating the degree of slope is easily done by the use of **equation 7**, and in general it is desirable to limit the slope at the minimum point of $R1$ so it does not exceed approximately 2.5 times the slope where $K$ is equal to 1.5. These calculations are not difficult to make, but they are time consuming and cumbersome, and can be avoided by using the value of $K$ at the minimum point of $R1$ as the sole criterion. When this value of $K$ is not less than 0.80, the slope ratio will closely approximate 2.5 as a maximum.

The practical significance of permitting a slope ratio of 2.5 is its indication that the adjustment of the loading control will be 2.5 times as broad at the minimum point of $R1$ as it is for the values of $R1$ in most of the *major-section*. Therefore a drastic change in tuning *feel* does not take place. Another point is the considerable response assymetry which the loading control exhibits as it is adjusted to either side of a particular point in the *minor-section*. The continually increasing slope of the $X_{C1}$ curve is the cause of this effect, and the degree of severity increases as lower values of $R1$ are used.

The following chart shows how rapidly the slope ratio increases as the value of $K$ is decreased; it may be used by the designer as an indication of what to expect if a $K$ of less than 0.80 is used.

| $K$ | slope ratio |
|---|---|
| 0.80 | 2.5 |
| 0.70 | 6.6 |
| 0.65 | 11.5 |
| 0.60 | 23.5 |
| 0.55 | 67.4 |

When the value of $K$ at the minimum point of $R1$ is close to but less than 0.80, the designer may find it helpful to know what this out-of-limit condition means in terms of $R1$, i.e., what is the value of $R1$ when $K = 0.80$, and how does it compare with the desired minimum value of $R1$? This information is obtained by simply multiplying the value of $R1B$ by 0.80 and comparing the result with the desired minimum value of $R1$.

From the foregoing it is apparent that it may be necessary to modify some design specifications to permit the use of the pi network by itself. In addition, some specifications will be such that they are realizable by only the addition of supplemental impedance transformation.

Since output loads below 32 ohms are in the *minor-section* of the prototype network the operation of this network with the loads defined in the analysis would require its classification as a compromise design. At $f_o$, or the frequency where $X_L = 200\ ohms$, $K = 0.81$ when the load is 26 ohms. If $f_o = 3.74\ MHz$, and a load of 26 ohms is maintained, $K$ will become 0.93 and 0.71 at 3.5 MHz and 4.0 MHz, respectively. The value of $K$ at 4.0 MHz (0.71) is appreciably below the limit recommended above; in fact, $K$ is out of limit for all frequencies above 3.8 MHz.

## the pi-L network

From fig. 3 it is apparent that pi networks can be operated at high impedance transformation ratios if no limit is placed on the maximum value of $Q_o'$. However, the trade-off is excessive power loss in the network inductor, in even moderate power rf power amplifiers. Minimizing this source of power loss is the primary purpose for using supplemental impedance transformation; it also provides a match to a greater range of load impedances and avoids any need for operating the pi network in its *minor-section*.

The L-network is by far the best choice for providing the supplemental impedance transformation.* Its efficiency is superior to that of any other network, and only one inductor must be added to form the pi-L combination. **Equation 12** has been specially formulated for determining the required reactance of the L-section inductor for load vswr up to 2:1.

Reactance of L-section inductor =

$$X_L' = \sqrt{Z_o(2R1B = Z_o)} \qquad \textbf{(12)}$$

In this equation $Z_o$ is the characteristic impedance of the antenna system, and the value of $R1B$ is obtained from the graph of **fig. 3** and the known relationship between $R2$ and $T_r$. In this use of **fig. 3**, $Q_o'$ may be assigned any desired value as long as the resulting value of $R1B$ is not less than one-half the value of $Z_o$. Contrary to contemporary design practice, it is not sufficient to assign an arbitrary transformation ratio to the L-section. The value should be no more than is necessary to allow the pi section to operate in its *major-section*.

A step-by-step procedure for designing a pi-L network (such as might be used with a 4-1000A) is included here as a design example. For this example the following specifications were assumed:

*When supplemental impedance transformation is required it is obtained most efficiently by the use of an L-network. However, some additional switching is required for the L-section inductor, which in some cases may be impractical or undesirable. The aperiodic transformer (untuned) is an interesting alternative to the use of tuned networks because it can be designed to operate quite efficiently over the 3- to 30-MHz range. The transformers in mind are the types discussed by Jerry Sevick, W2FMI,[8] John Nagle, K4KJ,[9] and others which use iron or ferrite cores to obtain efficient operation over a wide frequency band.

No data is available on the use of such transformers in the manner suggested here, however, there appears to be no reason why it should not be practical so long as core saturation is avoided.

**table 2. Twelve complex impedance values which fall on a 2:1 vswr circle plotted on a Smith chart (fig.4), and the required pi-L network capacitance reactance values required to provide a match to a plate load resistance of 2500 ohms** (*series* $X_L = 73.94$ *ohms*)

| point on 2:1 vswr circle (fig. 4) | worst-case antenna load impedance | worst-case load plus +j73.94 ohms | equivalent parallel components | | required $X_{C1}$ | required $X_C$ of loading control, C1 | required $X_{C2}$ |
|---|---|---|---|---|---|---|---|
| | A | B | C | D | E | F | G |
| 1. | 26.00 j 0.0 | 26.00 +j 73.94 | 236.27 | +j 83.08 | 85.99 | 42.26 | 390.86 |
| 2. | 27.38 −j 10.27 | 27.38 +j 63.67 | 175.48 | +j 75.44 | 78.36 | 38.44 | 401.63 |
| 3. | 32.00 −j 20.78 | 32.00 +j 53.16 | 120.30 | +j 72.42 | 70.98 | 35.85 | 414.52 |
| 4. | 41.60 −j 31.20 | 41.60 +j 42.74 | 85.51 | +j 83.23 | 67.40 | 37.24 | 425.79 |
| 5. | 59.43 −j 38.60 | 59.43 +j 35.34 | 80.44 | +j 135.28 | 67.23 | 44.91 | 427.82 |
| 6. | 86.60 −j 32.47 | 86.60 +j 41.47 | 106.45 | +j 222.31 | 69.28 | 52.82 | 418.57 |
| 7. | 104.00 j 0.0 | 104.00 +j 73.94 | 156.57 | +j 220.22 | 75.83 | 56.41 | 405.60 |
| 8. | 86.60 +j 32.47 | 86.60 +j 106.41 | 217.36 | +j 176.88 | 83.71 | 56.82 | 393.94 |
| 9. | 59.43 +j 38.60 | 59.43 +j 112.54 | 272.55 | +j 143.92 | 90.16 | 55.43 | 385.45 |
| 10. | 41.60 +j 31.20 | 41.60 +j 105.14 | 307.33 | +j 121.60 | 93.89 | 52.98 | 380.77 |
| 11. | 32.00 +j 20.78 | 32.00 +j 94.72 | 312.40 | +j 105.53 | 94.42 | 49.83 | 380.12 |
| 12. | 27.38 +j 10.27 | 27.38 +j 84.21 | 286.39 | +j 93.11 | 91.67 | 46.19 | 383.53 |

*Required plate load resistance = 5500 ohms*
$Q_0' = 14$
$Z_0' = 52$
*Maximum vswr = 2:1*

The design proceeds in the following steps:

**1**. Referring to fig. **3**, determine the value of $T_r$ when $Q_0' = 14$. In this case $T_r = 70$.

**2**. Compute $R1B$ by dividing $R2$ by $T_r$. $R1B = 5500/70 = 78.57$ *ohms*.

**3**. Compute $X_L$ by substituting $T_r$ and $R2$ in **equation 9**. $X_L = 464.83$ *ohms*.

**4**. Compute $X_L'$ by substituting $R1B$ and $Z_0$ in **equation 12**. $X_L' = 73.94$ *ohms*.

**5**. Add $X_L'$ to each of the worst-case load impedances (from previous example) and tabulate as in **table 2**.

**6**. Compute equivalent parallel resistance, and equivalent parallel reactance for each value of load impedance and tabulate as in **table 2**.

**7**. Compute required values of $X_{C1}$ and $X_{C2}$ for each value of equivalent parallel resistance. Use **equations 4** and **5** with *464.83 ohms* substituted for $X_L$.

**8**. Compute required reactance of loading control capacitor by combining the reactances obtained in **step 6** with the values of $X_{C1}$ obtained in **step 7**.

**9**. Repeat **steps 5** through **8** with $X_L$ and $X_L'$ appropriately modified to obtain the band-edge values.

## summary

The problems encountered in applying pi network design **eq. 1**, **2**, and **3** are due to the fact that these equations should be used only for fixed parameter networks, *i.e.* $R1$ and $R2$ should not be changed from the design values, and the operating frequency should not be varied. Nevertheless it has become common practice to use these equations for the design of pi networks for use in amateur equipment, and to assume that there is adequate tuning flexibility for the application. Unfortunately you can't depend on this assumption, especially when you want to operate with a vswr as great as 2:1.

It is interesting to note that a pi network, identical to the prototype network, can be designed using **eqs. 1**, **2**, and **3**, and a value of **14** for $Q$. From the previous discussion it is known that the tuning range of this pi network will meet the desired vswr conditions. Next consider a case similar to the prototype network except that R2 is changed from **2500** to **3500** ohms. **Eqs. 2** and **3** will yield a value of **273.39** ohms for $X_L$, but the designer using this procedure is not normally aware of the fact that this value of $X_L$ results in a value of **42.71** ohms for $R1B$ so the tuning range of the network is limited.

On the other hand, the designer using the procedure recommended in this article is immediately aware of the problem in the first two steps of the design procedure; Allowing $Q_0'$ to be **15**, **fig. 3** shows that $T_r$ is about **83**, and $R1B$ will therefore be **42.17** ohms (3500 ÷ **83**). Thus it is immediately known that this network cannot be expected to work well with low values of antenna load impedance.

At this point the designer may elect to use one of the four options in **step 5** of the step-by-step design procedure. These alternatives give direction to the course of action to be taken when the initial network design falls short of the desired objectives. Regardless of which alternative is ultimately chosen, the designer's choice will be based upon specific knowledge of how the pi network will perform in practice.

Designers who must frequently design pi networks will find the discussion in the *appendix* especially useful. It shows how universal design curves can be developed from the two new equations which have been derived from **equations 4** and **5**. The new equations are unique in that the only independent variables are $T_r$ and $K$, where $K$ is $R1$ normalized with respect to $R1B$. Families of curves drawn from these equations can be made to cover any desired range of pi networks.

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## appendix

### universal design curves

Curves of $X_{C1}$ and $X_{C2}$ are shown in **fig.** 2, and although the contours of these curves are typical of all pi networks, they apply specifically only to the prototype network. Universal design curves can be plotted to cover any desired range of possible networks, but before this can be done **equations 4** and **5** must be manipulated to obtain equations which are expressed only in terms of $T_r$ and $K$, where $K$ is the factor obtained when $R1$ is normalized with respect to $R1B$

$$K = \frac{R1}{R1B} \qquad \text{(I)}$$

Thus **equation 4** converts to

$$F_1 = \frac{1}{1 + \frac{\sqrt{T_r(K - 0.50)}}{K}} \qquad \text{(II)}$$

and **equation 5** converts to

$$F_2 = \frac{1}{1 + \sqrt{\frac{K - 0.50}{T_r}}} \qquad \text{(III)}$$

where $F_1$ and $F_2$ are proportionality factors which must be multiplied by $X_L$ to obtain specific values of $X_{C1}$ and $X_{C2}$. Therefore

$$X_{C1} = F_1(X_L) \qquad \text{(IV)}$$

$$X_{C2} = F_2(X_L) \qquad \text{(V)}$$

$T_r$ has already been defined as the ratio of $R2$ to $R1B$

$$T_r = \frac{R2}{R1B} \qquad \text{(VI)}$$

Since the factors $T_r$ and $K$ represent ratios of real values, but have no dimensions of their own, they meet the essential requirements for complete generality. **Equations II** and **III** are also completely general, and can be used in the plotting of universal curves.

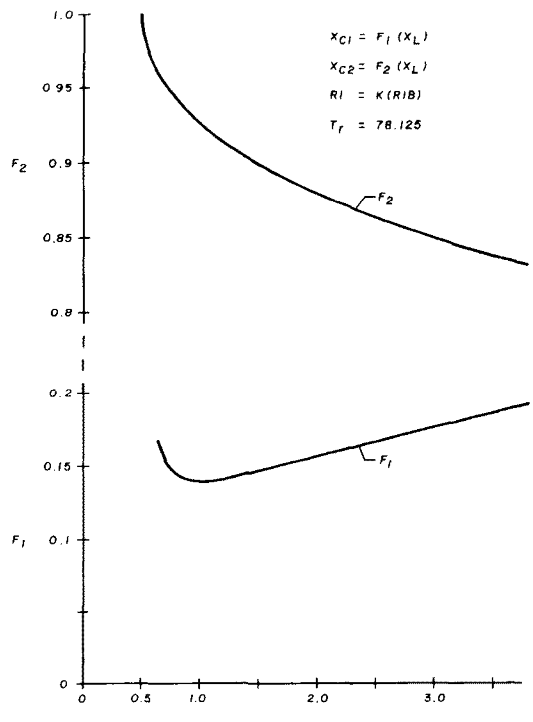
Curves plotted in Cartesian coordinates can accommodate only one variable in addition to the dependent variable so it is necessary to plot a family of curves for both **equations II** and **III**. Individual curves of a family should be based upon a selected value of $T_r$, and $K$ from 0.70 to 4.0 to cover the normal ranges of $R1$. Adjacent curve spacings depend on the selected values of $T_r$ and should be chosen to minimize interpolation errors.

In using the curves the designer reads the coordinate values of $F_1$ and $F_2$ for a particular value of $K$, and then converts them to values of $X_{C1}$ and $X_{C2}$ in accordance with **equations IV** and **V**. The value of $X_L$ in these equations may be obtained by using **equation 9** or the convenient form

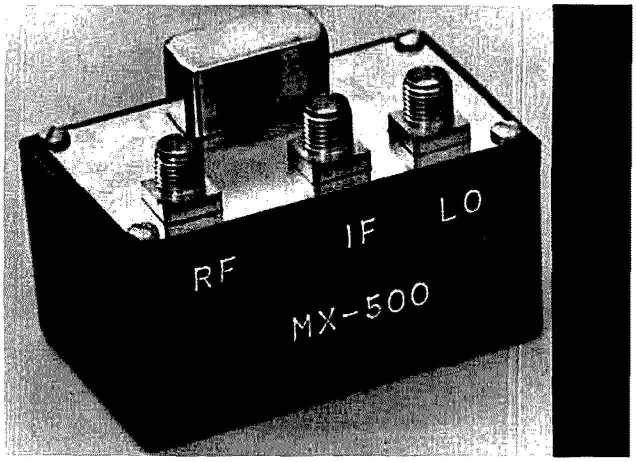
$$X_L = \frac{R2}{\sqrt{2T_r}} \qquad \text{(VII)}$$

The value of $K$ represents a particular value of $R1$ as given by

$$R1 = K(R1B) \qquad \text{(VIII)}$$




In the above graph the curves of **fig. 2** have been redrawn in accordance with **equations II** and **III**. Normally the curves would be based upon integer values of $T_r$; however, in the prototype network $T_r$ has the fractional value of 78.125 and the curves of the above graph are therefore based on this value. These curves can be used for all pi networks where $T_r = 78.125$.




# circuit packaging for uhf double-balanced mixers

Versatile PC board for dual-inline packaged double-balanced mixer modules provides flexible operation from dc to 500 MHz

**The use of double-balanced mixers** in transmit and receive converters has been explored in numerous magazine articles.[1-5] The construction of mixers for vhf and uhf service is a relatively straightforward matter, and consists of a quad of matched Schottky-barrier diodes in a ring or bridge arrangement and two wideband toroidal transformers.[6] However, with the cost of commercial double-balanced mixer modules now less than $10, it hardly pays for the experimenter to build his own.

I was first introduced to commercial double-balanced mixers by Joe Reisert, W1JR, who showed me how to use a dual-inline packaged mixer in a 432-MHz converter.[7] He later published information on a DIP pinout which is now used by many manufacturers for dc-500 MHz mixers.[8] (see **table 1**). The standardization of flatpack microwave mixers allowed me to develop a universal PC layout for use at 1296 MHz.[9] In this article I will present a similar PC layout for use with any of the uhf mixer modules listed in **table 1**.

## mixer circuits

The double-balanced mixer circuit shown in **fig. 1** is based upon an article by Reisert.[8] Except for different pin-numbering schemes used by various manufacturers, **fig. 1** is an accurate representation of all the mixers in **table 1**.

Note that two pins (3 and 7 in **fig. 1**) are connected together to form the i-f port. In some (but not all) mixers, these pins are tied together internally. To build a circuit board which is compatible with all the

By **H. Paul Shuch, WA6UAM**, Microcomm, 14908 Sandy Lane, San Jose, California 95124

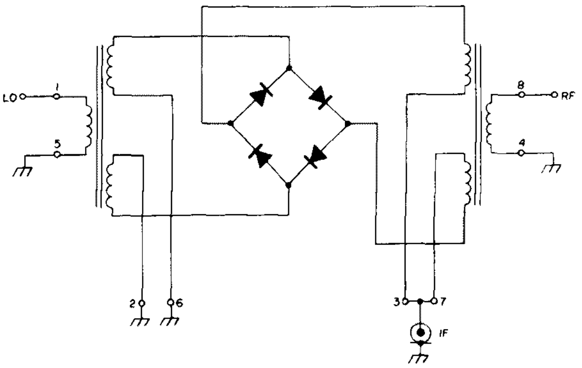



**fig. 1. Practically all uhf double-balanced mixer modules use the circuit shown hore, consisting of four hot-carrier diodes and input and output transformers. Some units are grounded internally, as discussed in the text.**

mixers in **table 1**, there must be a circuit trace between these pins. Also, in some mixers, the ground points (shown as pins 2, 4, 5, and 6 in **fig. 1**) are internally connected to the mixer case; in others they are not. Therefore, the PC board *must* provide for external grounding of these pins.

Of course, in mixers which are not internally grounded, it would be possible to connect pins 2 and 6 together for use as the i-f port, and ground pins 3 and 7. Similarly, if pin 1 were grounded, pin 5 could serve as the LO port. The same possibilities hold for

**table 1. Pin compatible dual-inline package double-balanced mixers. This list is not complete but gives an indication of the wide variety of pin-compatible mixers available on the market.**

| type | frequency range (MHz) | isolation (dB) | price (approx) |
|---|---|---|---|
| Anzac MD108 | 5 - 500 | 25 | $ 8 |
| Anzac MD109 | 0.2 - 200 | 25 | 17 |
| Anzac MD142 | 10 - 1000 | 20 | 55 |
| Mini-Circuits SBL-1 | 1 - 500 | 30 | 4 |
| Mini-Circuits SRA-1 | 0.5 - 500 | 30 | 10 |
| Mini-Circuits SRA-5 | 5 - 1500 | 25 | 30 |
| Merrimac 117A | 0.5 - 500 | 30 | 10 |
| Merrimac DMS-2-200 | 1 - 400 | 30 | 25 |
| Cimarron CM-1 | 5 - 500 | 25 | 6 |
| Cimarron CM-2 | 5 - 1200 | 20 | 15 |
| Summit 769E | 5 - 500 | 30 | 25 |
| Summit 761 | 3 - 1000 | 35 | 40 |
| Watkins-Johnson M6E | 5 - 500 | 30 | 37 |

Anzac Electronics, 39 Green Street, Waltham, Massachusetts 02154
Mini-Circuits Lab, 837-843 Utica Avenue, Brooklyn, New York 11203
Merrimac Industries, 41 Fairfield Place, West Caldwell, New Jersey 07006
Cimarron Division, Vari-L Company, 3883 Monaco Parkway, Denver, Colorado 80207
Summit Engineering, Post Office Box 938, Bozeman, Montana 59715
Watkins-Johnson Company, 3333 Hillview Avenue, Palo Alto, California 94304

pins 4 and 8 at the rf port. For that matter, in any double balanced mixer the rf and LO ports are completely interchangeable. However, since grounding is to be provided on the PC board, an internally grounded mixer will require a particular orientation on the board (more on this later).

When the mixer has been properly grounded, and the i-f pins have been tied together, the three mixer ports can be connected to coaxial connectors. To minimize impedance discontinuities it is advisable to use 50-ohm microstrip transmission lines when interfacing the mixer to coaxial connectors.

## circuit board

**Fig. 2** is a full-sized layout of a circuit board which will accommodate any of the mixers listed in **table 1**. As with all microstripline circuits, it is etched on one side of double-sided printed circuit material. The other side remains unetched and serves as a groundplane. The dimensions of the circuit were chosen to provide a good impedance match to 50 ohms at all

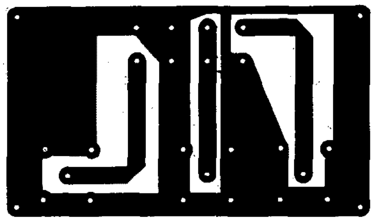


**fig. 2. Full-size printed-circuit board for dual-inline packaged double-balanced mixer modules.**

ports (1/16 inch or 1.5mm thick fiberglass-epoxy circuit board). At the operating frequencies of these mixers the thickness of the copper cladding is of little consequence; I used 2-ounce copper (about 2.8 mils or 70 microns thick) with no observed difficulties in either performance or etching.

The active pins of the mixer, as well as the center pins of the three coaxial connectors, must all be isolated from ground. This can be easily accomplished by using a 1/8 inch (3mm) twist drill as a countersink to remove the groundplane metallization from around the active pins as shown in **fig. 3**. Drilling instructions for the etched board are also shown in **fig. 3**. Note that all mixer and coax connector pins require no. 56 (1.2mm) clearance holes, while the no. 42 (2.4mm) mounting holes in the corner of the board easily accommodate no. 4 (M3) mounting hardware. This circuit board is designed to be used as the top cover of a Pomona Electronics 2417 die-cast aluminum box. This enclosure provides excellent shielding as well as an attractive appearance, as seen in the photograph.

## coaxial connectors

High quality coaxial connectors are recommended to minimize impedance mismatches at uhf. I have had excellent results with type SMA connectors, a military designation standing for Sub-Miniature, Type A. The SMA connector is a gold-plated precision threaded unit, with a 3mm reference plane dimension. They were originally developed by the Omni-Spectra Company under their brand designation OSM, a name by which they are often referred regardless of the manufacturer. SMA-compatible connectors have become a standard catalog item of numerous companies; the ones I use are E. F. Johnson JCM series. Their female chassis connector (part no. 142-2098-001) costs less than $3.00 and operates well into the microwave region. For interconnection to other modules, you will want to make up a few jumper cables. These can be fitted with E. F. Johnson 142-0261-001 plugs which mate well with the female chassis connector and cost about the same.

## mixer pinouts

If you examine the data sheets of the mixers listed in **table 1**, you can be easily misled into believing that the devices have different pinouts. This is because the manufacturers use different pin numbering sequences, as shown in **fig. 4**. Fortunately, all suppliers mark pin 1 in some way, usually by providing an odd-colored glass bead at the seal. Regardless of the pin numbering scheme, the internal mixer configuration is basically the same. Thus it's possible to install all these mixers on the same circuit board.

Installation of the mixer on the board is straightforward. With the board oriented as shown in **fig. 3** (groundplane side up, coax connectors toward you), position the mixer so pin 1 (the one with the odd-colored bead) is away from you and toward the right. The pins will fall readily into place, and can be soldered to the microstriplines on the opposite side of the board.

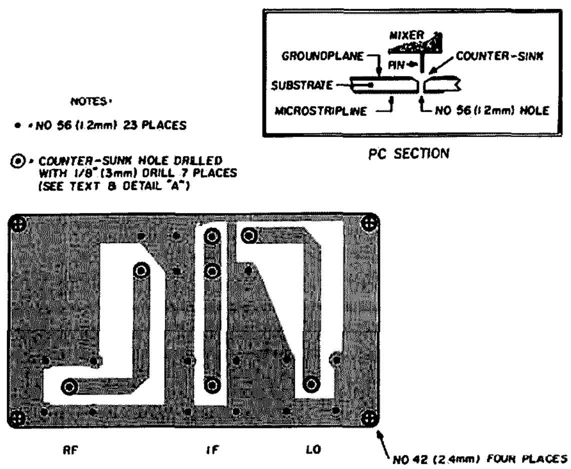



**fig. 3. Drilling instructions for the double-balanced mixer circuit board. Indicated holes must be countersunk to eliminate short circuits to ground.**

When installing the coax connectors, run a bead of solder around the connector body on the groundplane side of the board; then solder the four ground pins on the microstripline side. This provides "through-the-board" grounding of the applicable mixer pins.

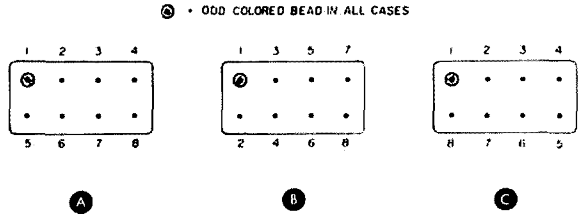



**fig. 4. Pin numbering arrangements used by various manufacturers of double-balanced mixer modules.**

## parts availability

Most of the manufacturers listed in **table 1** will sell their mixers directly to the individual experimenter in small quantities; a few may require that orders be placed through a regional representative. Circuit boards can be etched from the artwork in **fig. 2**.*

*Etched, drilled, and plated circuit boards are available for $4.50 from Microcomm, 14908 Sandy Lane, San Jose, California 95124, postpaid.

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**ham radio**

# the frequency counter as a synthesizer

Low-cost ICs provide a method of converting your counter to a frequency synthesizer with 10-Hz readout

**If you're planning to build** a frequency counter here's an inexpensive method for using your counter as a frequency synthesizer with increments of 10 Hz. This frequency synthesizer can be used with any VFO provided an afc input (tuning diode) is available or such modification is possible.

Because of the availability of low-cost integrated circuits, digital frequency counters for receivers, transceivers, and signal generators have become very popular.

In the case of receivers and transceivers in which the frequency being counted is different from the receiving or transmitting frequency, a programmable divider (up-down counter) can be used for i-f preset.

**Fig. 1** shows a block diagram of a frequency counter with an internal resolution of 10 Hz with the 100-Hz digit displayed. The 10-Hz digit is not displayed because the error of ± 1 digit is avoided as well as the annoying appearance of this digit at random intervals. The last (10-Hz) digit determines the highest frequency of operation. Using a low-cost Fairchild type F10010 BCD decade counter (**fig. 2**) **2**), operation to 200 MHz is provided.

Various methods have been developed in the past that are suitable for frequency synthesis. The most elegant, accurate, and expensive method is to use a sample-and-hold discriminator, which samples the frequency at 10-Hz intervals. This method is expensive. It requires lots of pulse-shaping circuitry and therefore is vulnerable to thermal drift. Probably the simplest way of doing this, but not the most accurate, is a method in which a statistical error is determined. **Fig. 2** shows such a circuit.

## average circuit

The disadvantage of the circuit in **fig. 2** is that, because of the average reading, an error of 1 digit is possible (10 Hz). Also two power supply voltages are required. The digital-analog converter converts the ECL pulses from the F10010 counter into a dc voltage, which is then integrated and, through a darlington emitter follower, is available as the afc voltage. While tuning, the numbers 4, 5, 6, or 7 appear at the output of the units decade counter. This means a high output level (L in ECL level) at the $Q_3$ output of the F10010 counter. Therefore, CR24 ceases to conduct. Q31, Q32 conduct, and the voltage across C55 becomes ± 1.3 V. Thus the output voltage becomes − 2.5 V and remains constant during tuning. Once tuning is completed the output will be different from 4, 5, 6, or 7. In the event that the reading is 1, 2, or 3 the voltage across capacitor C54 will decrease. If the result is 8, 9, or 0 the voltage will increase. Ideally, the average from counting should be 0.5; however, because of component tolerances the last digit could be ± 1.

The output of this circuit must be connected to the oscillator, and an increasing voltage will cause a higher frequency. Once the loop is closed, the final result will become 0.5 as a statistical average; therefore the frequency will be stabilized with respect to the counter reference frequency.

## comparator circuit

Because the circuit of **fig. 2** is somewhat limited by drift and component tolerances, a slightly more complex circuit offers the best solution. It uses the 4-bit comparator type 7485. When extremely low power consumption is vital, the National CMOS 74C85 can be used. **Fig. 3** shows the circuit, which is capable of determining counting differences between 0 and 15, so any odd number between 1 and

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16 can be used. Sometimes using a divide-by-16 rather than a divide-by-10 mode results in the upper cutoff frequency being slightly higher.

The 4-bit comparator receives information from a 7485, which is used as a register and from the actual count of the first known displayed digit. The comparator has two pulse-train outputs that must be integrated and which are used to charge or discharge integrating capacitors. Because of the logic decisions that have to be made, this circuit can't be used below 0 and above 15 because the comparator will not respond to the upper and lower limits. Therefore additional circuitry is required to avoid misreading.

The desired values are now divided into three categories,

**A**. Small values (0-3).

**B**. Medium values (4-11).

**C**. Large values (12-15).

The actual (count) values are divided into two categories,

**D**. Small values (0-7).

**E**. Large values (8-15).

In cases **A** and **B** the desired value is a small number and the count value is a large number; in cases **C** and **D** the desired value is a large number and the count value is a small number. The connection between the comparator output and the charge-pump inputs in this case must be exchanged. In all other cases no change is needed.

Assume the case in which the desired value is 15

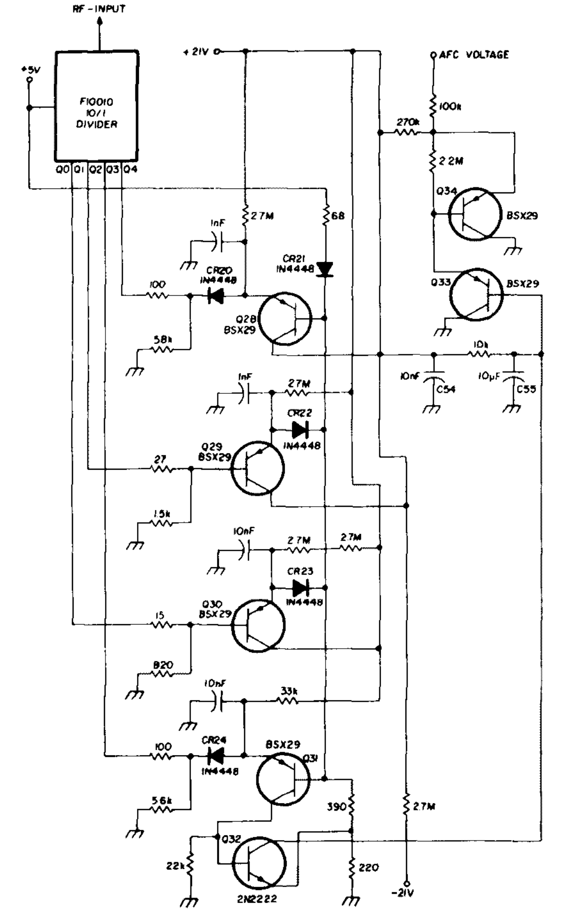



**fig. 2. Averaging circuit to obtain an afc voltage from the first ECL counter in fig. 1.**

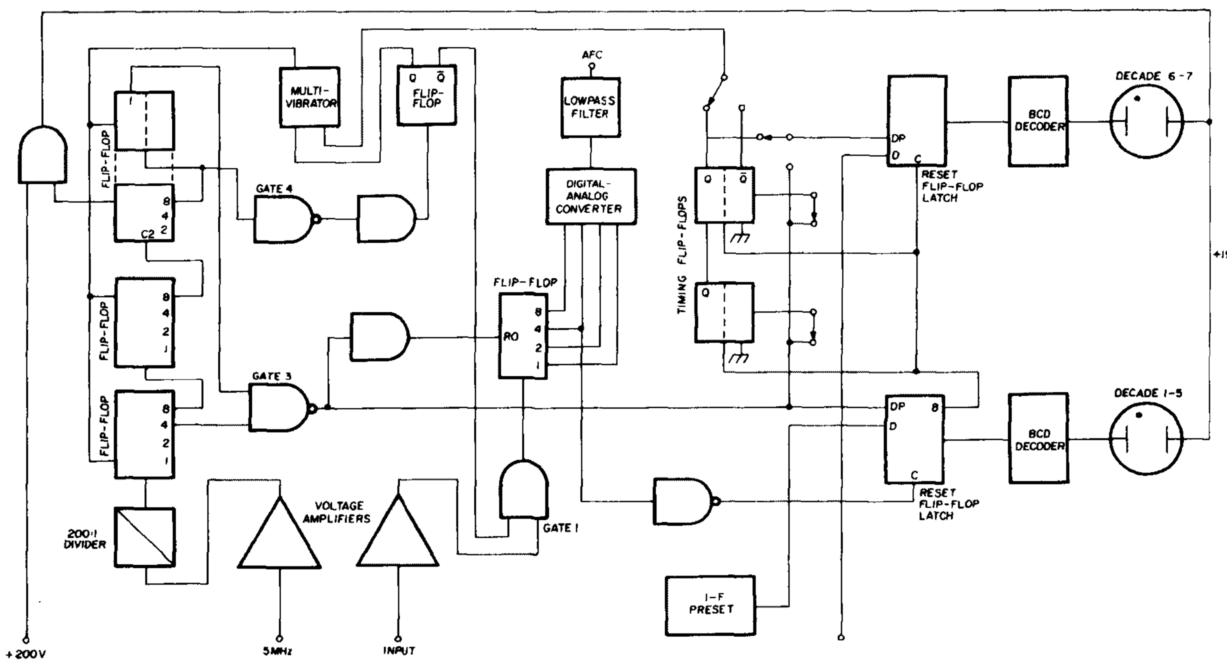



**fig. 1. Block diagram of a frequency counter-synthesizer with i-f preset capabilities and nixie display.**

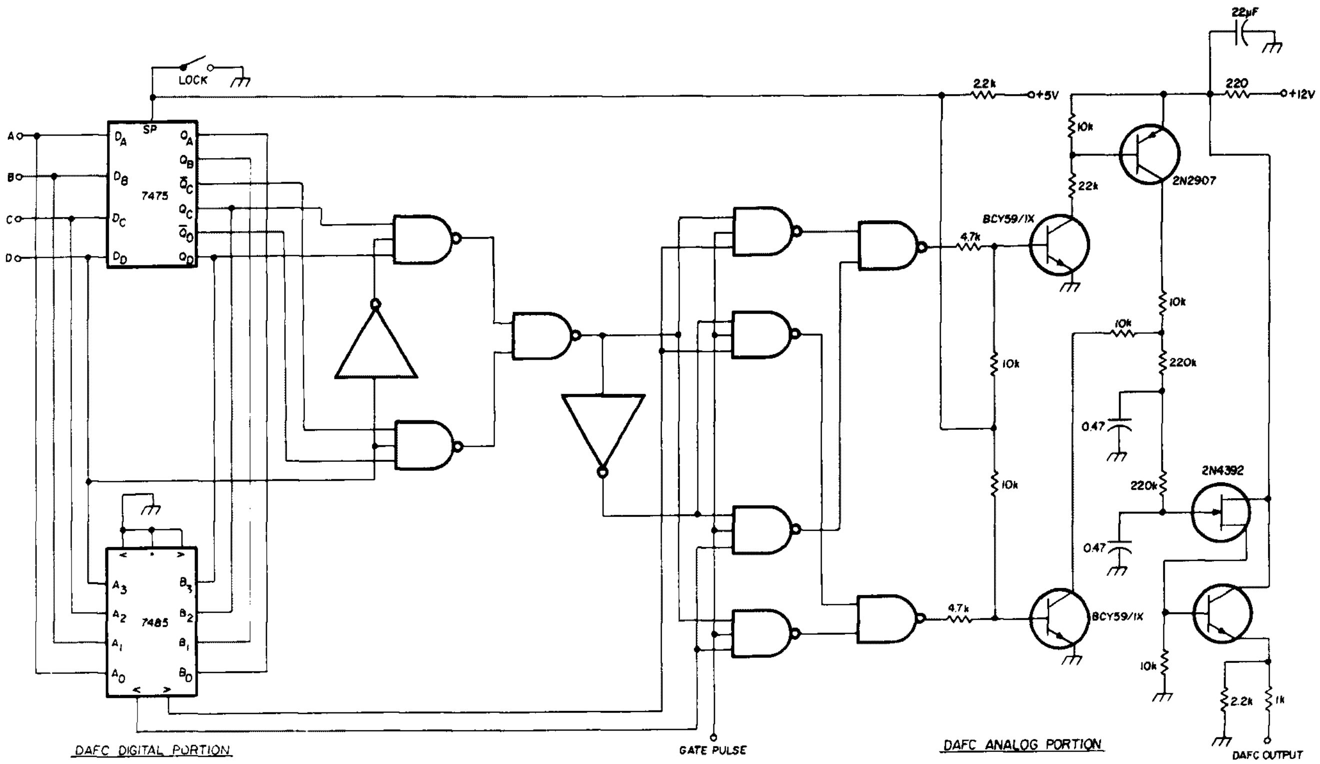



fig. 3. Comparator or digital automatic frequency correction circuit (DAFC). Inputs A, B, C, D must be connected to the level-shifter outputs of the ECL A, B, C, D decade counters/4-bit binary counters as shown in fig. 4.

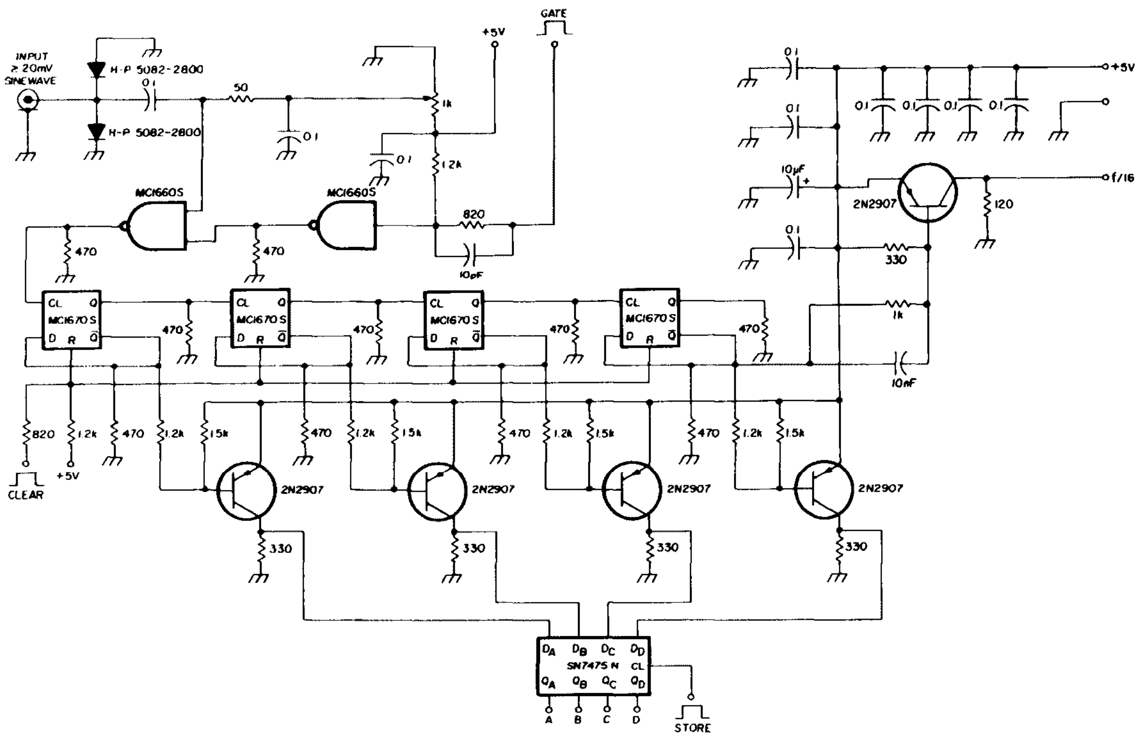



fig. 4. Input stage for a 350-MHz frequency counter and level shifters for the TTL stages. Input threshhold level can be adjusted by a 1-k pot.

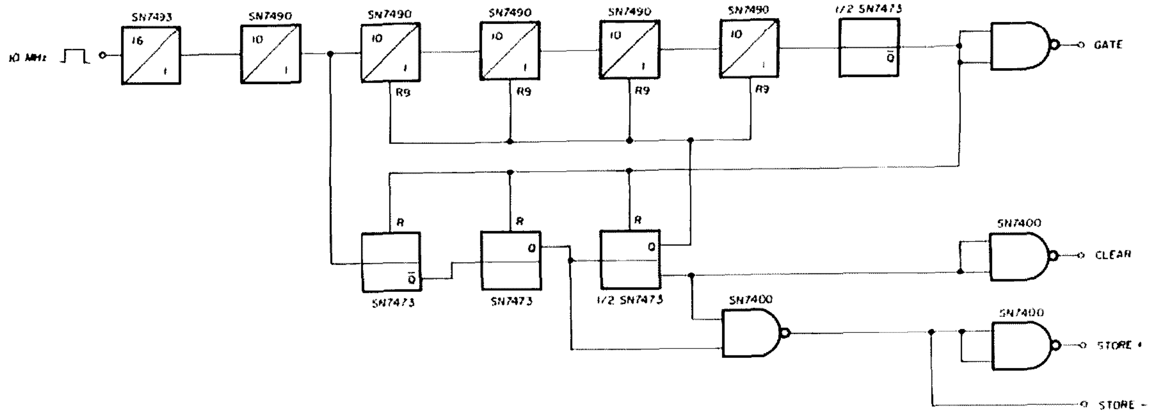
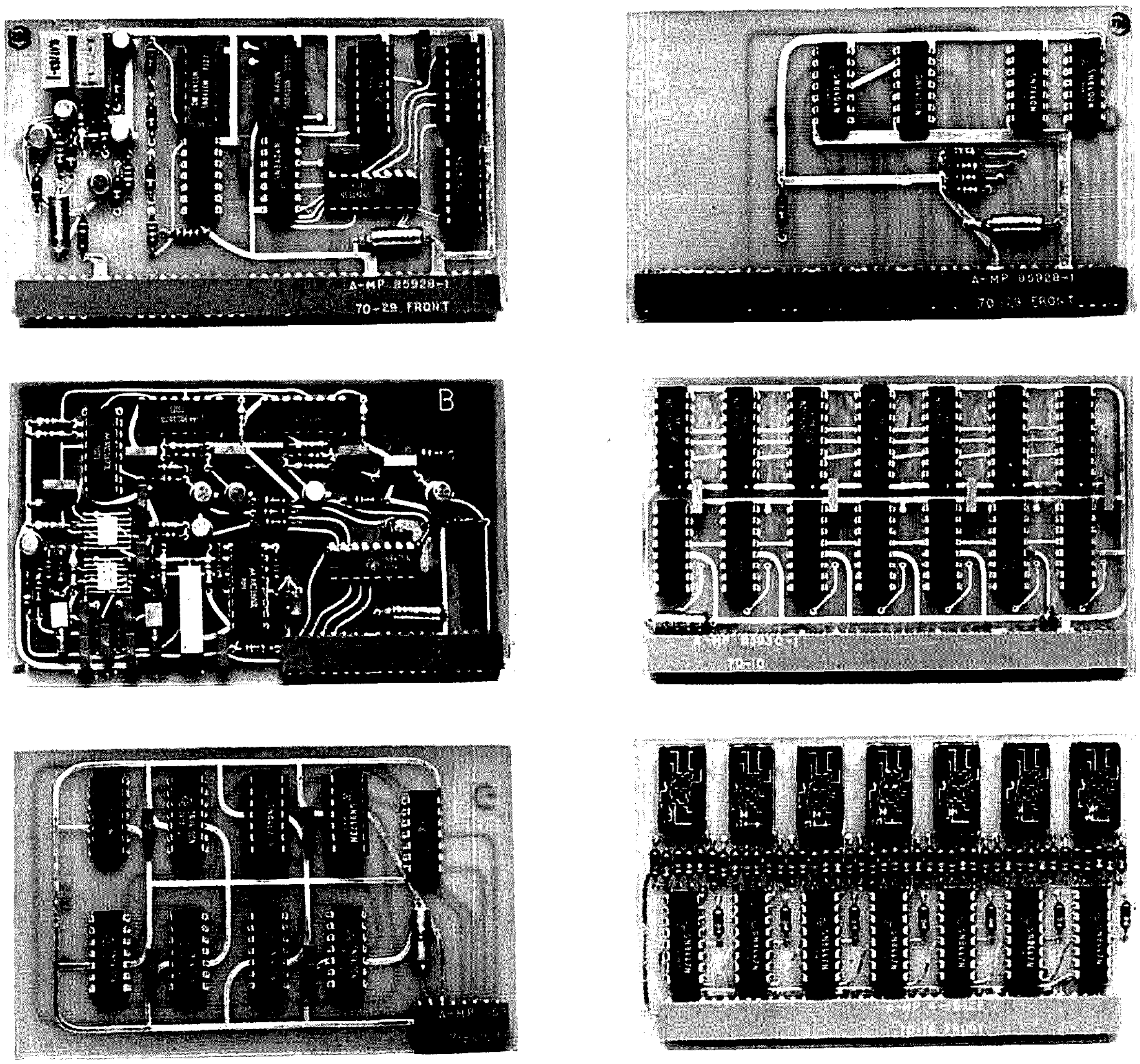



**fig. 5. A recommended time base/generator.**



Photographs of the circuit boards developed for the counter. Starting from the upper left and working clockwise: PC board using a similar circuit as in fig. 3. To achieve a digital sweep capability, the 7475 was replaced by an 8493. Programming card for the i-f preset. Up/down counters and latches. BCD/7-segment decoder and MAN-1 displays. Time base divider/generator. Input stage for a 350-MHz counter/synthesizer for 100/16-Hz locking steps. The schematic is shown in fig. 4.

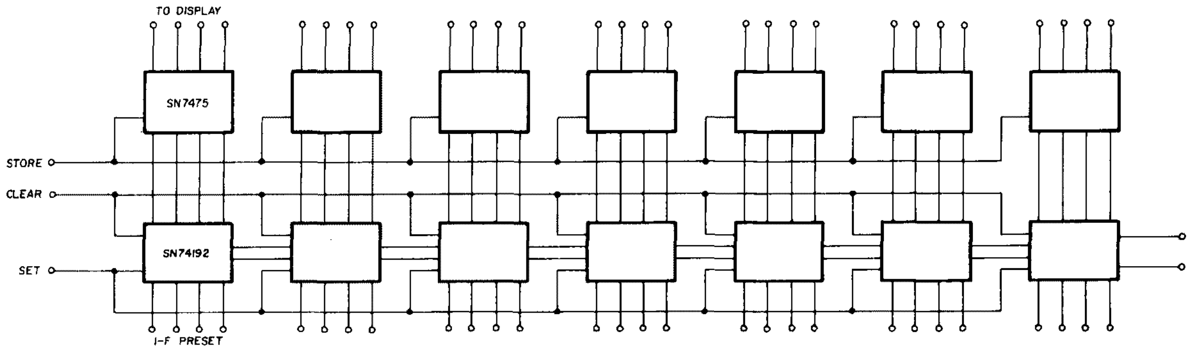



**fig. 6. Programmable dividers and latches.**

and the oscillator drifts to a higher frequency. The count value should read zero. In this case the comparator reads an actual value smaller than the desired value; however, input and output have been exchanged with the aid of an additional gate, since 15 is a large number and zero is a small number (cases **C** and **D**). As mentioned before, the circuit has been designed to read values between zero and 15, so it will also work using a decade counter. In this case,

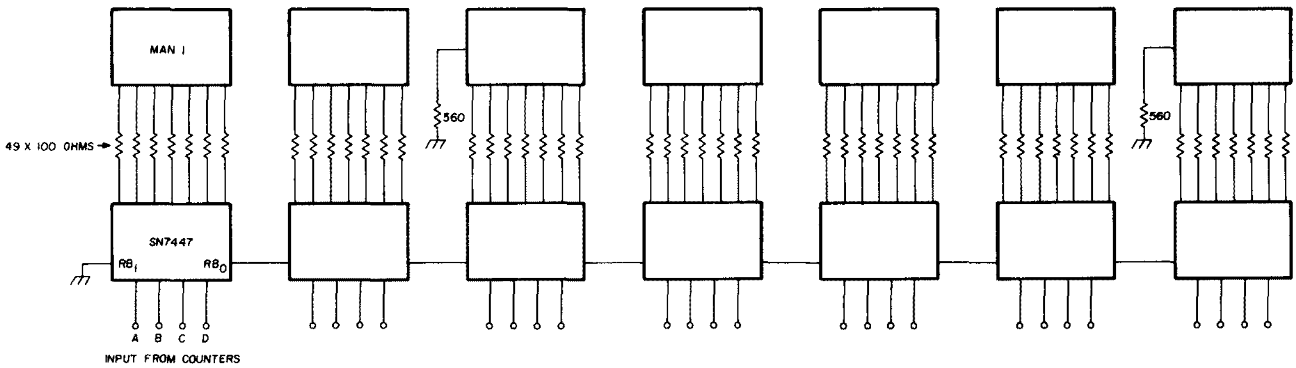



**fig. 7. BCD 7-segment decoders and MAN-1 display.**

however, actual count and preset value can only be between zero and 10. If type 74192 programmable counters are used instead of the 7490 decade counter, i-fs can be preprogrammed and therefore actual readings can be obtained.

## 350-MHz system

For even higher resolution such as 5.25 Hz (100 Hz divided by 16), **fig. 4** shows the input portion of a counter for the divide-by-16 requirement. It works to about 350 MHz. The circuit is basically as that described above using the F10010. **Fig. 5** shows the time-base generator circuit. Note the first 1/16 divider, which compensates for the 1/16 divider of the counter input. **Fig. 6** shows the programmable dividers and latches. Frequency stepping in 5.25-Hz increments is possible by feeding suitable pulses into the up-down counters. The BCD decoders and MAN-1 displays are shown in **fig. 7**. The polarity of the diode and the pulling range are important for stability considerations.

This circuit provides good stability over an ±8-kHz pulling range and, depending on the gate frequency, compensates for a drift of 25-100 Hz. During the initial switch-on period, most transceivers built by radio amateurs drift 1 or 2 kHz, so this circuit compensates for this effect.

## concluding remarks

No PC-board layouts are available for the counter; however, wiring can be accomplished easily by using perf boards. The input gates and flip-flops shown can be substituted with low-cost MECL-II ICs, such as the MC1007 or MC1027. However, this will reduce the maximum input frequency to around 100 MHz.

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**ham radio**

# an amateur hydroelectric station

Amateurs are known for their ingenuity — this article describes a primary electrical system powered by a water wheel

By I. L. McNally, K6WX, 26119 Fairlane Drive, Sun City, California 92381

**This is a report on the development** of an amateur hydroelectric system, which is located in the wilderness of the California mountains. The development began in 1932 when Arch Warnock, W6GQL, and his wife, Marian, W6HQV, acquired a mountain cabin located far from telephone or power lines. Arch and Marian decided they needed a way to keep in touch with home, and this is when I was asked to help with the hydroelectric project.

This article presents the development of the hydroelectric power system as it evolved from the first primative system to that used at present, which includes features such as pushbutton control and direct-reading circuits for line-voltage frequency. Information is also included on how to determine water flow rate and electrical and hardware requirements for duplicating this system. The information serves as a baseline for developing a system of even higher sophistication.

The first power source (in 1932) was an auto battery, which energized a dynamotor. The dynamotor supplied plate voltage for a single type 210 tube crystal oscillator. The inconvenience of hauling the battery down the mountain for charging prompted the idea of using water power for battery charging from a nearby mountain stream. Thus the amateur hydroelectric system described here was born.

Our first water wheel was built in 1933 — a low-pressure, high-volume affair 2 feet (0.6m) in diameter. It was made of sheet metal and was mounted in an open 4 x 4 foot (1.2x1.2m) wooden frame. An old Ford car wheel with a flat rim served as the driving wheel for an arrangement of V pulleys on a counter shaft. Sufficient speed was thus obtained to drive a 15-volt dc Dodge car generator. Seventy feet (21.3m) of 8-inch (20cm) irrigation pipe carried a 5-foot (1.5m) head of water along the natural gradient of the stream-bed. The water was applied undershot to the wheel, and the open construction caused a lot of spray. A few well-placed shields made it possible to approach the generator for servicing.

This assembly underwent minor modifications during the next several years. Several different generators, including a modified rotary converter for ac, were used but the maximum power generated never exceeded about 200 watts. The first hydroelectric plant was washed away in a flood that occurred in 1938.

The second water wheel, which was built in 1938, was considerably improved. It had been suggested by two local FCC inspectors, John and James Homsy. The rear end of an old Essex auto was used because it had a high gear ratio. One axle was removed and the housing was cut back to the center

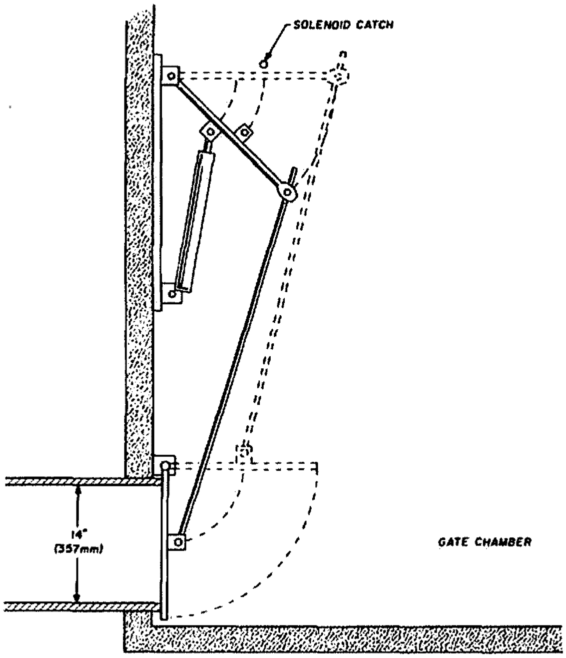
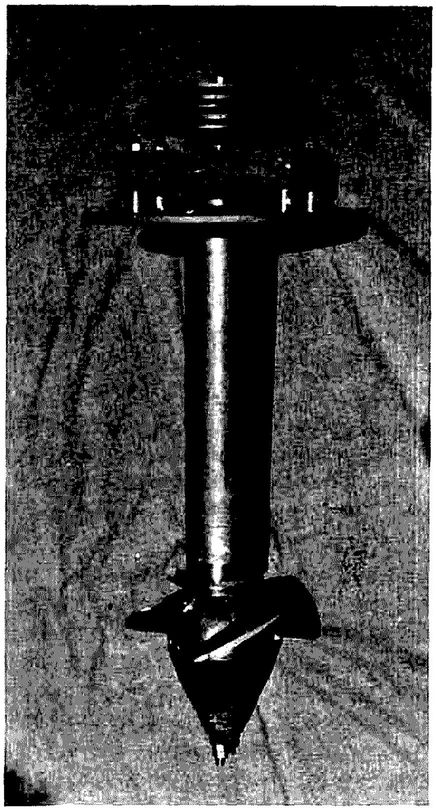



**fig. 1. Hydraulic gats liftar for controlling water into the turbine. A hydraulic piston and lever system powered by a 12-volt-dc hydraulic pump operates the mechanism.**



**Photograph of the complete impeller assembly.**

and sealed. Both brake drums were placed back-to-back on the remaining axle. Around the circumference ten 10-inch (25.4cm) concave tank tops were welded radially to the drums at equal intervals. The wheel, which was 3 feet (0.9m) in diameter, was operated undershot (to maximize the head with a 4-inch (10.2cm) nozzel). A 14-inch (36cm) wooden V

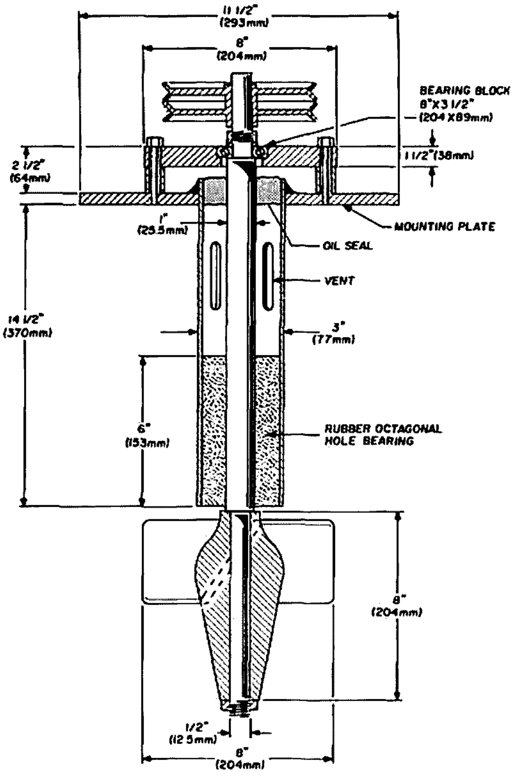



**fig. 2. Leffel turbine impeller. System is designed to deliver 2 hp (1492 W) at 1200 rpm from a 10-foot (3m) head of water supplying 155 cfm (4.3cmm).**

pully replaced the universal joint. The 10-foot (3m) head of water was carried to the wheel through a 10-inch (25cm) diameter concrete pipe, 60 feet (18.3m) long. To provide protection from future floods, the assembly was housed in a concrete shelter, which was excavated into the stream bank.

The third system, built in 1945, evolved into the present system. The vertical housing was made by using a 14-inch (36cm) section of an old 12-inch (30cm) diameter hot water tank. The upper end was closed with a mounting plate that had an 8½-inch (22cm) hole. The lower end was tapered to an 8-inch (20cm) diameter throat into which a boat propeller was suspended. The propeller suspension shaft was housed in a 3-inch (7.6cm) OD pipe, which was

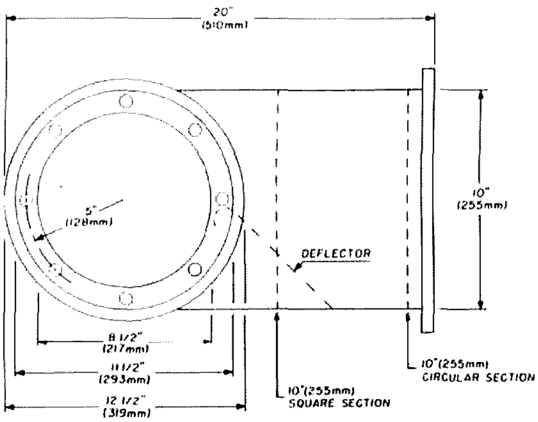
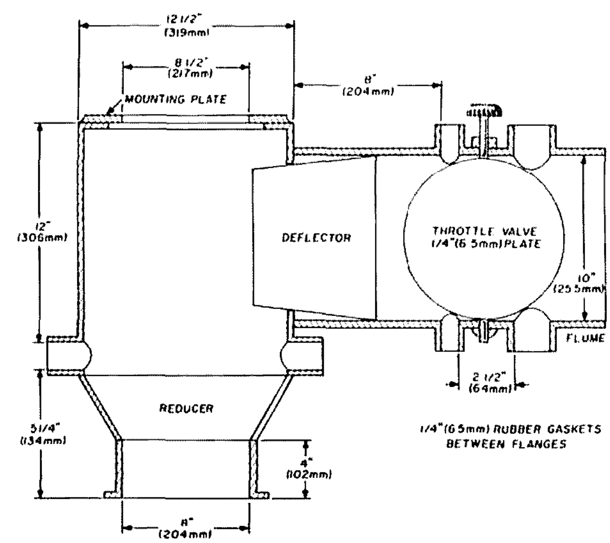



**fig. 3, Turbine chamber detail. The throttle valve is controlled from several remote locations. Deflector plate imparts a spinning motion to the water as it enters the turbine housing.**




welded to the top cover plate. Water brought in at a right angle to the upper chamber drove the propeller in conventional turbine fashion.

The fourth system, which was started in 1958, uses an 8-inch (20cm) Hoppes-type HL impeller supplied by the Leffel Company, which is designed to deliver 2 hp (1492 W) at 1200 rpm from a 10-foot (3m) head of water at 155 cfm (4.3cmm). The turbine impeller is made of bronze and consists of four curved rectangular blades mounted around a tapered hub. The blades are attached to the hub at a 45-degree angle and are machined to occupy the inside of the throat with a 1/16-inch (1.5mm) clearance. The lower end of the shaft, which is in the water, rotates in a hard-rubber boat-propeller shaft bearing. The upper

**table 1. data for determining the amount of water flow across a weir.**

| water depth, D, in. | (mm) | water flow, F, cfm | (cmm) |
|---|---|---|---|
| 1 | (25.5) | 0.4 | (0.0004) |
| 2 | (51) | 1.14 | (0.0013) |
| 3 | (76.5) | 2.00 | (0.0022) |
| 4 | (102) | 3.22 | (0.0036) |
| 5 | (127.5) | 4.50 | (0.0049) |
| 6 | (153) | 5.90 | (0.0065) |
| 7 | (178.5) | 7.44 | (0.0083) |
| 8 | (204) | 9.10 | (0.0101) |
| 9 | (229.5) | 10.86 | (0.0121) |
| 10 | (255) | 12.71 | (0.0141) |

bearing, which is housed in an aluminum-alloy block, is from a Ford Fairlane automobile wheel.

## design

The first step in planning a hydroelectric system is to determine the quantity and head of water available. The approximate flow of a stream may be determined by measuring the average stream depth, $D$, average stream width, $W$, and average distance of water flow per minute, $L$. The formula is

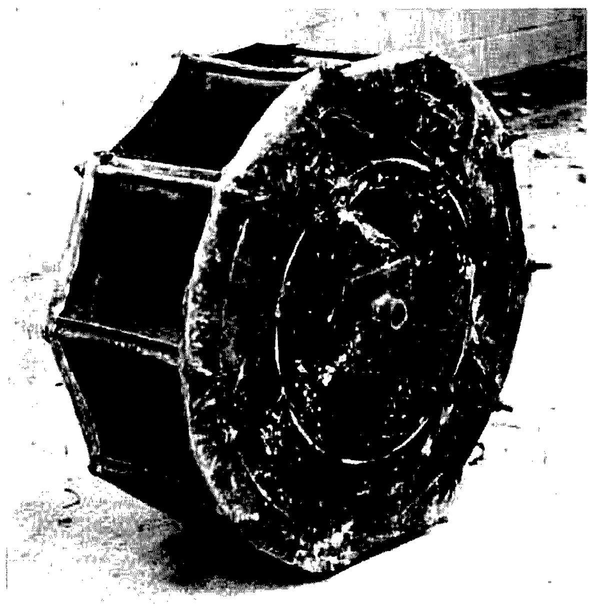
$$F = 0.8WDL \tag{1}$$

where

$F$ = water flow rate
$W$ = stream width
$D$ = water depth
$L$ = water flow distance

The distance, $L$, can be determined by measuring the movement of a half-submerged float in the stream for a given period of time.

**Example:** Water depth is 0.75 foot (0.23m), stream width is 5 feet (1.5m), and float movement is 10 feet



**First water wheel, which was made of sheet-metal.**

(3.0m) in 12 seconds, or $L = 50$ feet (15.2m). The flow rate, $F$, in this case is

$$F = 0.8 \times 0.75 \times 5 \times 50 = 150 \text{ cubic feet per minute (cfm)}$$

In metric terms the water flow rate is

$$F = 0.8 \times 0.23 \times 1.5 \times 15.2 = 4.2 \text{ cubic meters per minute (cmm)}$$

A more accurate method is to use a weir. A weir, consisting of a plank placed across the stream, must be constructed as the top section of a dam. The plank has a notch cut in it that is about six times as wide as the depth of water flowing over it. The notch is beveled on all sides with a sharp edge on the downstream side. A stake is placed about 3 feet (0.9m) upstream with its top level with the bottom of the weir notch to measure water depth, $D$. **Table 1** gives data for determining water flow, $F$, in terms of weir width for water depth up to 10-inches (25.4cm). For example:

6 inches of water over a 3 foot wide weir:

$$36 \times 5.90 = 212.4 \text{ cfm}$$

The metric equivalent is:

$$918 \times 0.0065 = 5.97 \text{cmm}$$

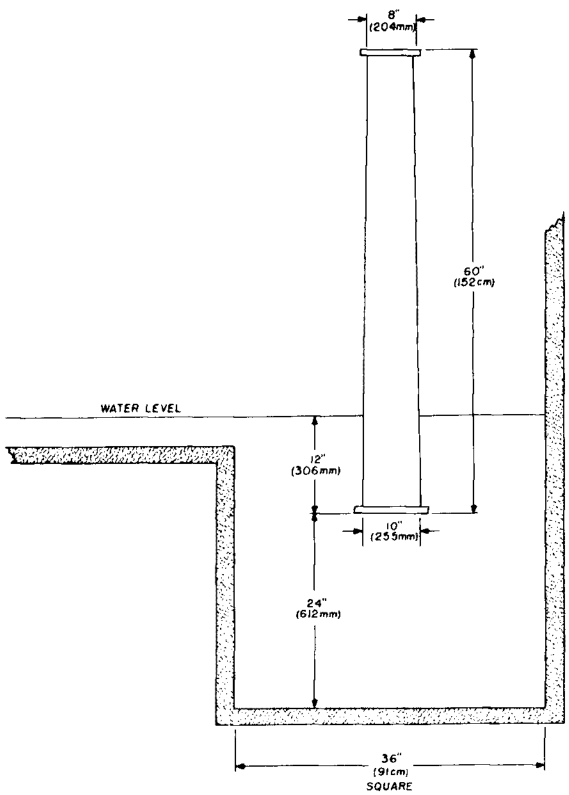



**fig. 4. Discharge pipe (or draft tube) assembly and exhaust chamber, which are essential for good turbine efficiency.**

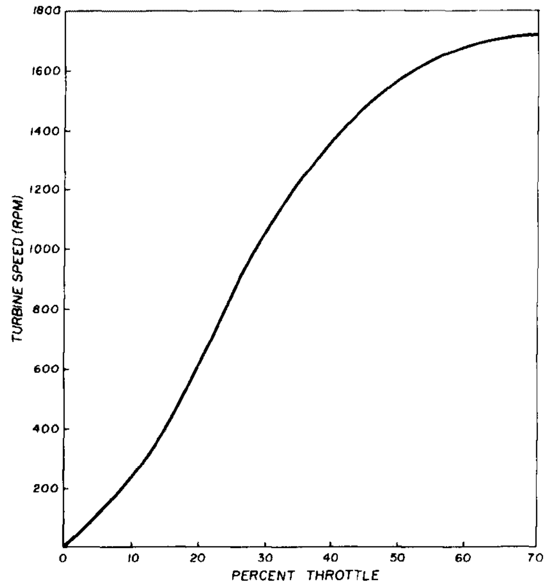



**fig. 5. Turbine speed as a function of percent throttle opening with 500 cfm (14cmm) of water flow.**

The next measurement to be made is the water height at the turbine. In the present system the dam and stream gradient gave a 10-foot (3m) head of water with 60 feet (18m) of pipe. All but 4 feet (1.2m) of this flume is 10-inches (25.4cm) in diameter. The flume was made as smooth as possible with a minimum of discontinuities.

**Intake chamber.** The intake chamber is located about 4 feet (1.2m) upstream from the dam. It is protected by a trash screen about 4 feet (1.2m) high, which is made of 5/16-inch (8mm) vertical rods spaced 1-inch (25mm) apart. The distance from the trash screen to the 14-inch (36cm) opening into the flume pipe is 3½ feet (1m).

The first 4 feet (1.2m) of the flume pipe is 14-inch (36cm) in diameter. This pipe is attached to a 3-foot (0.9m) sheet-metal cone, which tapers from 14 to 10 inches (36cm to 25cm). The remainder of the 60 feet (18m) of flume pipe is 10-inches (25cm) in diameter.

A surge chamber, which is located about 10 feet (3m) from the turbine is also used. It consists of 6 feet (1.8m) of 10-inch (25cm) diameter pipe mounted vertically on a concrete pad. A 4-inch (10cm) opening in the top of the flume pipe allows water to rise in the vertical surge chamber.

**Gate lifter.** When the turbine is not operating, a horizontally hinged ½-inch (12.5mm) circular iron plate closes the intake opening (**fig. 1**). The plate is raised to the open position by a hydraulic piston and system of levers and a 12-volt dc airplane hydraulic pump located in the turbine house. (A stand-

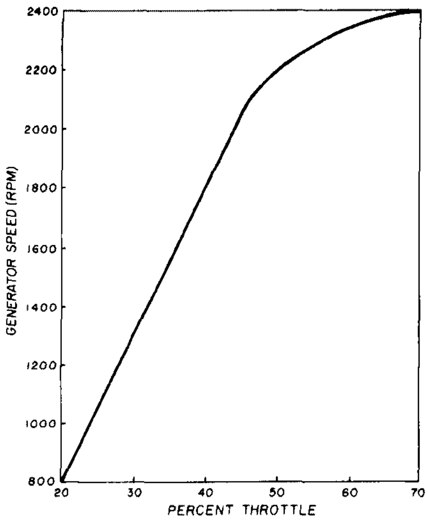



**fig. 6. Generator speed as a function of percent throttle opening with 500 cfm (14cmm) of water flow. Pulley ratios were adjusted to give 1800 rpm (60 Hz line frequency) at 40 percent throttle to provide a margin when water supply is low.**

by, manually-operated pump also opens the gate.) The gate is held in the open position by a solenoid-operated catch. To close the gate, the solenoid-operated catch is released, while hydraulic fluid is bled into the cylinder. Gravity and the force of water close the gate tightly against the opening.

**Propeller.** The Leffel turbine impeller (**fig. 2**) revolves counterclockwise, viewed from above, in a 4-inch (10cm) deep throat 8-1/16 inch (21cm) in diameter. The shaft, which is 1 inch (25.5mm) in diameter, is housed in a 3-inch (77mm) OD pipe, which has a B.F. Goodrich no. RFG 16P hard-rubber bearing at the lower end. A Ford Fairlane wheel bearing, part no. BCA RW-207-CCR, which is mounted in an aluminum block, is located 1-inch (25.5mm) above the turbine housing. An oil seal, National no. 5441130, at the top of the shaft prevents water leakage.

**Turbine chamber.** This assembly is shown in (**fig. 3**). A circular metal-plate butterfly valve is mounted in a section of the 10-inch (25cm) pipe just before the pipe enters the turbine chamber. The lever arm of this valve is actuated by a 12-volt dc reversible motor-driven screwjack, which is controlled from several remote locations. A deflector plate is placed at an angle to give the water a spinning motion as it enters the turbine housing.

**Draft tube.** After the water has passed the impeller, it enters the discharge pipe, which is called a draft tube (**fig. 4**). This tube, which is a 5-foot (1.5m) vertical pipe, expands from a diameter of 8-inches (20cm) at the top to 10-inches (25.5cm) at the lower end. A concrete discharge chamber, measuring 3 x 3 x 3 feet (0.9x0.9x0.9m), provides water overflow into a tailrace. The draft tube and discharge chamber, with the dimensions shown, are essential for good turbine efficiency.

**Impeller efficiency.** To determine the electrical power to be expected it's necessary to know the efficiency of the turbine impeller. From the data furnished with the Leffel impeller, the efficiency is 68 per cent. With 200 cfm (5.7cmm) of water available

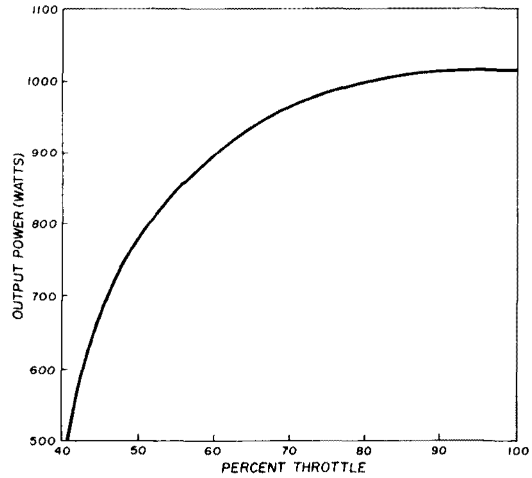



**fig. 7. Power output as a function of percent throttle opening with 200 cfm (5.7cmm) of water flow available.**

and a 10-foot (3m) head, 100-percent efficiency would give a power of

$$P = \frac{62.4 \times F \times H \times 746}{33{,}000} \qquad (2)$$

where

$P$ = power (watts)
$F$ = water flow (cfm)
$H$ = water head (feet)

Thus,

$$P = \frac{62.4 \times 200 \times 10 \times 746}{33{,}000} = 2821 \text{ watts}$$

Substituting metric equivalents for water flow, $F$, and water head, $H$:

$$P = \frac{62.4 \times 5.6 \times 3 \times 746}{277} = 2823 \text{ watts}$$

Under these conditions, about 1 kW will be delivered for an overall efficiency of approximately 35 percent.

**System data.** Data was plotted to show turbine

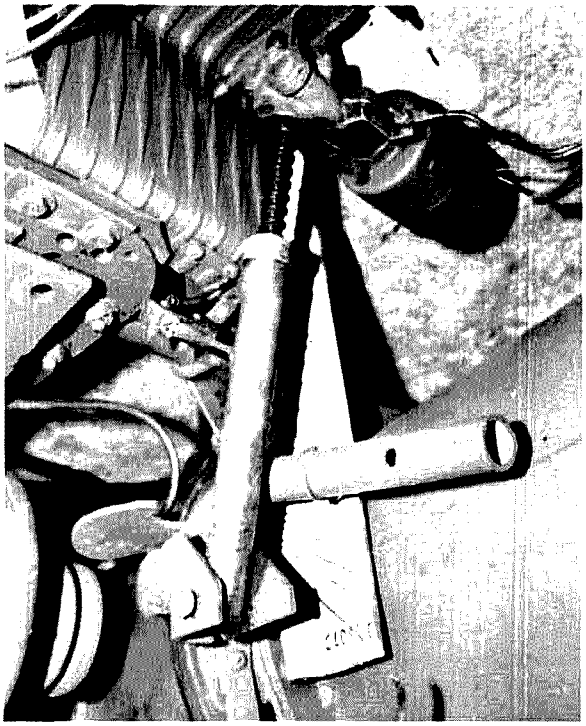
speed as a function of throttle opening (**fig. 5**). Data was also plotted to show generator speed with pulley ratios to give 1800 rpm (60 Hz) at 40-percent throttle. (See **fig. 6**). This allows a margin during times when the water supply is low. A third graph (**fig. 7**) shows power output under these conditions.

## control circuits

Most of the control circuits perform simple switching functions. However, a couple of ideas are offered for those interested in monitoring line frequency and building automatic control circuits for the dc-motor relays.

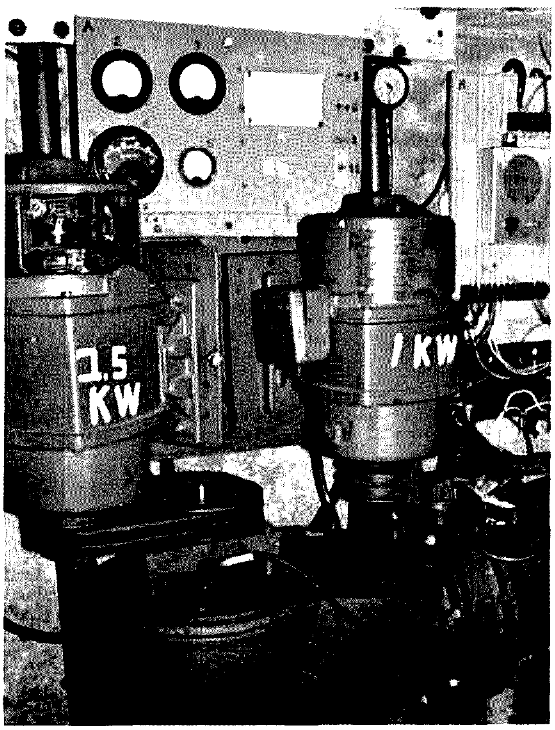
**Line-frequency metering.** Most surplus meters for measuring low-frequency ac voltage cover 60 ± 2 Hz. To obtain a greater frequency range a parallel LC network was used, which resonates at about 25 Hz. The linear portion of the resonance curve extends to 40-80 Hz, which gives a direct reading of line frequency.

**Automatic controls.** Manual pushbutton control is used for the reversible 12-volt dc motor, which moves the throttle-valve arm to the open or closed position. An automatic circuit energizes these motor-control relays by using an old Jewel ac meter and three photoelectric cells. The three photocells are located under the meter pointer at the 100-, 115-, and



Throttle valve showing screwjack, lever arm, and limit switches.

130-volt positions. When the meter pointer passes over the photocell at the low-voltage position, a relay opens the throttle. This relay is actuated by a transistor-operated sensitive relay and a latching relay. When the meter pointer passes the midpoint photocell the latching circuit opens, stopping the movement of the throttle throttle-valve arm. The same sequence occurs when the meter pointer passes the high-point photocell, which causes the throttle valve to close at midpoint. The throttle operates only when the load changes significantly.




Generator installation.

**Periperal circuits.** The ac generator and its standby circuit are mounted vertically on hinged plates, which allows either to be moved into position and driven by dual V belts. The throttle-valve control is also used in conjunction with a 5-kW variac. The original 15-volt Edison battery bank is still used to operate numerous controls as well as a 12-volt lighting system. A 50-ampere bridge rectifier allows the batteries to be charged at all times.

## acknowledgement

Many have contributed time and effort to this project, but two people deserve special thanks: Henry Backlund, a skilled machinist, and Jack Venturini, an expert mechanic and welder. Both worked full time on the amateur hydroelectric system.

**ham radio**

# how to design regulated power supplies

How to choose the correct values for components in your regulated supply — a typical 5-volt supply is used as a design example

**Probably one of** the most overlooked areas of electronic circuit design is the power supply. Try finding information on how to design a power supply and you'll probably find either a circuit that doesn't fill your needs or a design procedure (if you can find one) so caged in higher mathematics as to make it undecipherable. The one mysterious area in power-supply design is the determination of capacitor size. How often have you asked yourself just what capacitor is needed for a power-supply circuit?

With modern solid-state equipment, a regulated power supply is a must. With the new "regulators on a chip," we need only a power transformer, a set of diodes, a capacitor, and the regulator chip. The question that still remains, however, is "what size of everything do I need?"

## determining component sizes

The problem is not as difficult as it first seems. Since diode and transformer sizes are most easily obtained, we'll look at them first. Having decided on a regulated supply, you obtain a regulator chip. These chips are available in a variety of voltages and current ratings and will probably be in a TO-3 case unless it's the adjustable-voltage variety. Having determined your requirements, you purchase a chip that has the desired voltage and current rating. A quick reference to the data sheet for the device (to find the maximum input voltage), and you're ready to design your circuit. Since the power-supply output is to be regulated, the voltage developed by the transformer can be anything that produces a dc voltage not greater than the regulator maximum input voltage, nor less than the minimum voltage required by the regulator. Expressed mathematically, this voltage is

$$E_{reg\,min} < E_c < E_{reg\,max} \tag{1}$$

where

$E_{reg\,min}$ = regulator minimum input voltage
$E_{reg\,max}$ = regulator maximum input voltage
$E_c$ = voltage across the filter capacitor

The reason for maintaining the input voltage above the minimum voltage is to establish a "guard voltage," which prevents the input voltage from decreasing below the regulator level. Any voltage between the minimum and maximum input voltage for the regulator will produce acceptable results. Convert this value to a transformer secondary voltage by the following formula:

For bridge and half-wave rectifiers,

$$E_{sec\,(rms)} = (0.707)\,E_c \tag{2}$$

For center-tapped transformers in full-wave supplies,

$$E_{sec\,(rms)} = (1.414)\,E_c \tag{3}$$

The transformer dc-current rating for full-wave supplies using center-tapped transformers may be obtained by

$$I_{dc} \geqslant I_L/1.6 \tag{4}$$

and for half-wave and bridge rectifiers

$$I_{dc} \geqslant I_L/0.8 \tag{5}$$

where $I_L$ is the amount of current in amperes that

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the supply provides. The symbol $\geqslant$ means that any value larger than that obtained will work. For example, if $I_L = 1$ ampere and $I_{dc} \geqslant I_L/0.8$, any transformer capable of supplying more than 1.25 amperes will work. The numbers 0.8 and 1.6 are safety factors to ensure that the transformer won't be underrated.

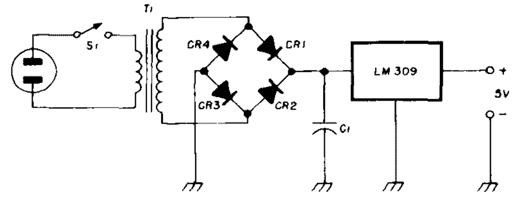


fig. 1. Bridge rectifier circuit used in the design example.

The current and peak-reverse voltage ratings can be obtained similarly. The peak-reverse voltage ratings for the diodes would be

$$E_{PIV(rated)} \geqslant \frac{E_{sec(peak)}}{0.8}$$

for all full-wave rectifiers and

$$E_{PIV(rated)} \geqslant \frac{2E_{sec(peak)}}{0.8}$$

for half-wave circuits.

The value $E_{sec\ (peak)}$ is obtained by multiplying the full secondary rms by 1.414.

The current ratings for the diodes are

$$I_{(average\ rated)} \geqslant I_L/1.6 \qquad (8)$$

for all full-wave rectifiers and

$$I_{(average\ rated)} \geqslant I_L/0.8 \qquad (9)$$

for half-wave rectifiers.

The remaining task is to determine capacitor size. Since the capacitor is rated according to capacitance and voltatge rating, a minimum value must be determined for each.

The capacitor will charge to the peak value of the voltage produced by the rectifiers, so the working voltage for the capacitors should be

$$WVDC \geqslant \frac{E_{sec\ (peak)}}{0.8} \qquad (10)$$

for non-center-tapped transformers, and

$$WVDC \geqslant \frac{E_{sec\ (peak)}}{1.6} \qquad (11)$$

for center-tapped transformers.

The value for $E_{sec\ (peak)}$ is found by multiplying the secondary rms voltage by **1.414**. For center-tapped transformers, use the entire secondary voltage.

The capacitor value can be found by

$$C \geqslant \left(\frac{I_L}{E_c}\right) X \qquad (12)$$

where $X$ is obtained from the accompanying tables. The value for $C$ is in farads and can be converted to microfarads by multiplying by $10^6$. $I_L$ is again the current that the supply is to provide, and $E_c$ is the voltage across the capacitor (obtained from the rectifier).

Although the tables are for 60-Hz line frequency, the values may be converted to any other line frequency (such as 400 Hz) by

$$X_{full\ wave} = \frac{120\ X_{FW}(from\ table)}{F} \qquad (13)$$

$$X_{half\ wave} = \frac{60\ X_{HW}(from\ table)}{F} \qquad (14)$$

where $F$ is the supply ripple frequency (Hz).

**Tables 1-3** provide a range of per cent ripple voltages between 0.1 and 32 per cent with corresponding factors for $X_{FW}$ and $F_{HW}$.

## design example

Now that we have the procedure down, let's see how it works. Assume we need a 5-volt supply capable of providing 1 ampere of current. Several regulator chips are available, such as the National Semiconductor LM-309. A quick reference to the

**table 1. Multiplying factors for use in eq. 12 to determine capacitor values (0.1 - 0.9 per cent ripple voltage).**

| per cent ripple voltage | $X_{(FW)}$ | $X_{(HW)}$ |
|---|---|---|
| 0.1 | 8.21 | 16.54 |
| 0.2 | 4.08 | 8.24 |
| 0.3 | 2.71 | 5.48 |
| 0.4 | 2.02 | 4.10 |
| 0.5 | 1.61 | 3.27 |
| 0.6 | 1.34 | 2.72 |
| 0.7 | 1.14 | 2.33 |
| 0.8 | 0.996 | 2.03 |
| 0.9 | 0.882 | 1.80 |

**table 2. Multiplying factors for use in eq. 12 to determine capacitor values (1-9 per cent ripple voltage).**

| per cent ripple voltage | $X_{(FW)}$ | $X_{(HW)}$ |
|---|---|---|
| 1 | 0.792 | 1.62 |
| 2 | 0.386 | 0.799 |
| 3 | 0.252 | 0.526 |
| 4 | 0.186 | 0.390 |
| 5 | 0.146 | 0.309 |
| 6 | 0.120 | 0.254 |
| 7 | 0.101 | 0.216 |
| 8 | 0.087 | 0.187 |
| 9 | 0.076 | 0.165 |

**table 3. Multiplying factors for use in eq.12 to determine capacitor values (10-32 per cent ripple voltage).**

| per cent ripple voltage | $X_{(FW)}$ | $X_{(HW)}$ |
|---|---|---|
| 10 | 0.0677 | 0.147 |
| 11 | 0.0607 | 0.132 |
| 12 | 0.0549 | 0.120 |
| 13 | 0.0500 | 0.110 |
| 14 | 0.0458 | 0.101 |
| 15 | 0.0422 | 0.0935 |
| 16 | 0.0390 | 0.0869 |
| 17 | 0.0363 | 0.0810 |
| 18 | 0.0338 | 0.0758 |
| 19 | 0.0317 | 0.0712 |
| 20 | 0.0297 | 0.0670 |
| 21 | 0.0279 | 0.0633 |
| 22 | 0.0263 | 0.0599 |
| 23 | 0.0249 | 0.0567 |
| 24 | 0.0235 | 0.0539 |
| 25 | 0.0223 | 0.0513 |
| 26 | 0.0212 | 0.0489 |
| 27 | 0.0201 | 0.0466 |
| 28 | 0.0192 | 0.0445 |
| 29 | 0.0183 | 0.0426 |
| 30 | 0.0174 | 0.0408 |
| 31 | 0.0167 | 0.0391 |
| 32 | 0.0160 | 0.0376 |

data sheet tells us that we should maintain the input voltage between 7 and 35 volts. If we use a bridge rectifier the circuit might look like **fig. 1**. If we convert the 7 and 35 volts to rms, then for the bridge rectifier these voltages would correspond to a transformer voltage (rms) range of 5 to 25 volts.

Although any transfomer within this range would work, the only common values are 6.3 and 12.6 volts rms. Either will work; however, the 12.6 volt rms transformer would probably be more economical since an increase in voltage decreases the filter-capacitor value.

Assume we'll use the 12.6 volt transformer, transformer. The current rating for the transformer would be

$$I_{dc} \geqslant \frac{1}{0.8} = 1.25 \; ampere$$

A transformer with a 1.5-ampere rating should do nicely.

The diodes should be capable of supplying

$$I_{dc} \geqslant \frac{1}{1.6} = 0.625 \; ampere$$

and have a reverse voltage rating of

$$E_{PIV} \geqslant \frac{(12.6)\,(1.414)}{0.8} = 23 \; volts$$

Diodes rated at 1 ampere at 50 volts should work well for CR1-CR4 in **fig. 1**; for example the type 1N4001.

Since the peak voltage across the capacitor will be about 18 volts, or 12.6 x 1.414, and any capacitor that will maintain the negative ripple peaks above 9 volts would be adequate. (Some ripple will appear at the output, however, and this value will be proportional to the ripple voltage across the filter capacitor.)

The ripple voltage, $E_r$, can be converted to per cent ripple by

$$per\ cent\ ripple = \frac{E_r}{E_{c(peak)}} \cdot (100) \qquad \textbf{(15)}$$

(See **fig. 2**) If $E_r$ is taken as maximum, the negative ripple peaks will be $V_I/0.8$ and the positive peaks will be

$$E_{c(peak)} - (12.6)\,(1.414) = 18 \; volts$$

(See **fig. 3**). The per cent ripple will then be

$$per\ cent\ ripple = \frac{E_{c(peak)} - 9}{E_{c(peak)}}$$

$$= \frac{18 - 9}{18}$$

$$= 50 \; per\ cent$$

Since the per cent ripple exceeds the values listed in the tables, we can use any value of capacitance determined from the tables, since these values will produce capacitor values with ripple less than 50 per cent. The voltage rating for the capacitor would be

$$WVDC \geqslant \frac{E_c}{0.8} = \frac{18}{0.8} = 23 \; volts$$

Choose capacitors with a maximum rating of at least 25 volts. Usually in a case such as this, the most difficult problem is determining the per cent ripple that can be tolerated. In the example above, the maximum ripple was 50 per cent. The amount of ripple we can allow in our power supply is anything between zero and 50 per cent. The only difference be-

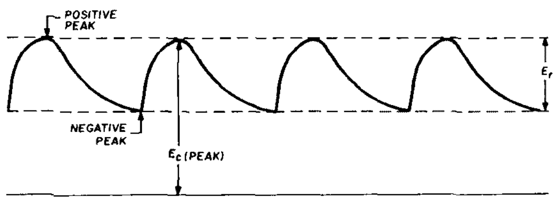


**fig. 2. Typical waveform showing ripple voltage, $E_r$, and voltage across filter capacitor, $E_c$.**

tween using a capacitor that will produce 50 per cent ripple and one that will produce 10 per cent ripple is the amount of ripple appearing across the regulator output.

The regulator acts as a filter (if $V_{in} < 7$ *volts*); however, some ripple will be conducted to the output. This value of ripple will be in the millivolt range in both extremes, with a higher value of output ripple for an input ripple of 50 per cent. The best sugges-

tion is to use a capacitor that produces the higher value of ripple when the supply is to power large-signal, low-gain circuits. Smaller values of ripple are dictated by higher gains and smaller signal levels. Ripple levels of 10 to 30 per cent will be adequate for all but the most stringent applications.

Increasing the ripple voltage above 30 per cent or decreasing the ripple by using capacitors that produce less than 10 per cent ripple, usually results in

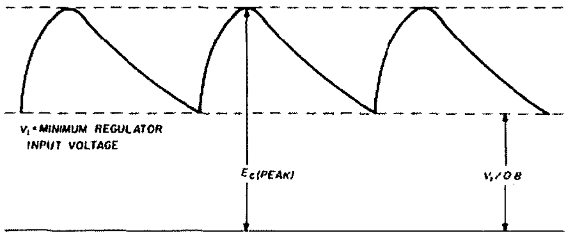


**fig. 3. Waveform showing relationship between positive and negative ripple-voltage peaks with respect to the minimum regulator input voltage.**

less-than-desired performance. The value of capacitance for 10 per cent ripple would be

$$C \geqslant \frac{XI_1}{E_c} = \frac{0.0677 \cdot 1}{18}$$

$$= 3.761\ (10^{-3})\ farads$$

$$= 3761\ microfarads$$

and the value for a capacitor producing 30 per cent ripple would be

$$C \geqslant \frac{(0.01744)\cdot(1)}{18} = 9.69\ (10^{-4})\ farad$$

$$= 969\ microfarads$$

A good range of values for $C$ would be about 1000-4000 microfarads.

## in summary

Components for the power supply producing 5-volts output and a maximum load current of one ampere would be a 12.6-volt transformer capable of providing 1.5 amperes, four diodes (such as 1N4001s), a capacitor between 1000 and 4000 microfarads with a voltage rating of 25 volts, and a 5-volt regulator. For noncritical applications I would use a 1000-microfarad capacitor, and where ripple may be a problem I would use a 3750 microfarad capacitor.

The procedure above is one I've found useful for the past several years. The ability to design and build one's own equipment is one of the most enjoyable facets of amateur radio. I hope you will find as much enjoyment in the design process as I have.

**ham radio**

# private call system for vhf fm

Tired of the blurps and bleeps while listening to your repeater? Try this tone encoder for private and noise-free monitoring

**There are many times** when you might like to monitor your local repeater while waiting for a friend to come on without having to listen to continual chatter or "kerchunking." The system described here is similar to the Motorola paging units wherein a tone is transmitted that disables receiver squelch. With the *Priva-Call* installed in your radio, all your friend has to do is transmit your particular (private) tone. The *Priva-Call* will sense this tone, connect the speaker to the rig, and a CALL light will illuminate. The system is self-contained and is simply plugged into your external speaker jack. A 10-ohm resistor provides a load to your radio at all times.

## circuit description

Audio from your receiver speaker is fed into pin 3 of U1, the NE-567V phase-locked loop (**fig. 1**). When a tone of the correct frequency is received, the voltage on pin 8 will drop from 4 to near zero volts. This causes Q1 to switch off, increasing its collector volt age from zero to 2 volts. When the gate voltage of Q2, the SCR, increases to a voltage higher than the cathode voltage, Q2 triggers on and closes relay K1, which turns on the speaker. The speaker will now remain on until the RESET switch is activated.

When you're finished answering your call, simply switch to RESET, then switch to ON. This will disconnect the speaker, and the *Priva-Call* will remain in the standby mode until the proper tone is again received.

## construction

The circuit was breadboarded and verified for operation using a plug-in type breadboard available through E. L. Company, Continental Specialties, Radio Shack, and others. This saves much time during the debugging stage.

After the circuit was working properly at minimum current drain, it was built onto a piece of Vector board 1 x 1½ inches (3x4cm) in size with hole spacing to match the NE567 PLL.*

Parts layout isn't critical, since the circuit operates at audio frequencies. By making the board size small enough, it was fitted inside my existing *Touch-Tone* box.† An alternative method would be to mount the board, a battery pack (9 or 12 volts), and a small speaker inside a separate box and mount this box next to your rig. The decoder plugs into the speaker jack and the tone generator taps into the microphone input, so the system can be changed easily from rig to rig without destroying the resale value.

The easiest way to construct the *Priva-Call* is to start at the left side of the circuit diagram and just begin wiring parts to the board in the order you get to them. Most component values are uncritical and can be obtained from advertisers in this magazine or through your local Radio Shack store.

R1 can be a small trimmer of 10k-100k ohms. R2 and R3 determine the phase-lock frequency and should total 1k-15k for most audio tones. The values I chose gave a frequency range from 830-1340 Hz. Reducing R2 to 1k and increasing R3 to 10k or 20k will give a much broader range. R4 can be 2k-22k. R5 can be 1k-10k. R6 and C5 provide a 1.5-second time delay to prevent false triggering. R8 can be 50 to 500 ohms; its purpose is to keep the gate current small. R9 and CR2 aren't really needed for proper operation and could be omitted. CR2 does provide a CALL in-

*Etched and drilled printed circuit boards are available from the author; instructions and parts layout are included. The price is $3.95 postpaid. Comments and questions are welcome, but please send a self-addressed stamped envelope for replies.

†Available through D.A.R.T., Post Office Box 201, Clawson, Michigan 48017. (Price as of 1975 was $6.75, postpaid.) For a picture and description, see *ham radio*, November 1973, page 67.

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dicator, though, and the combination of R9 and CR2 helps to reduce the current drain when the relay is on.

The value of R9 should be small enough so the relay will turn on and off reliably. Start with no resistor (also remove CR2 so it won't burn out), and get the circuit working first. Add R9 and CR2, then start increasing R9 to reduce current drain. By adding that one resistor, the current drain can be reduced from 40 to 25 mA — a valuable saving if your rig is battery operated.

C1 can be a 0.1 to 0.5$\mu$F disk ceramic. C2 determines the frequency and should be of good quality for stability. C3 should be a minimum of twice the value of C4. If C3 is too large, it will delay turn-on and turn-off of the NE567. C4 determines the bandwidth and should be 2.2$\mu$F. C6, C7, and C8 provide rf bypassing to prevent triggering by nearby transmitters.

For those who wish to experiment, the formula for frequency in this case is

$$F = \frac{1.1}{(R2 + R3) \cdot C2} \tag{1}$$

where $F$ is frequency (Hz), $R2$, $R3$ are in ohms, and $C2$ is in farads.

The formula for per cent bandwidth is

$$BW = 1070\,(V/F \times C4)^{1/2} \tag{2}$$

where $BW$ is bandwidth (%), $V$ is the input voltage (which equals 0.2 V) $C4$ is in $\mu$F, and $F$ is frequency (Hz).

## encoder circuits

Three simple ways can be used to provide the call ing tone; two are oscillator circuits and the third

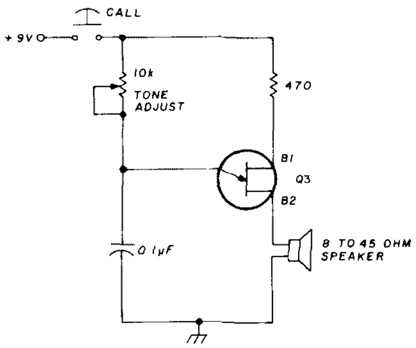


**fig. 2. Encoder system 2, which uses a unijunction transistor (UJT) oscillator. The speaker is held near your microphone to activate the system.**

uses a *Touch-Tone* pad. I used the tone pad since it provides a stable frequency by pushing adjacent buttons. See **table 1** for a listing of all the possible frequencies available.

System 2 is a UJT oscillator, which can be audio-coupled to your transmitter by holding the speaker next to the microphone (see **fig. 2**).

System 3 uses a phase-locked loop IC as the tone generator and is directly connected to the microphone input, **fig. 3**. Also, bear in mind that these are not the only circuits that will work for generating audio tones. For example, many circuits have been devised using the NE555 timer IC.[1]

## system hookup

The *Priva-Call* system is designed to fit between the speaker jack on your transceiver and an external speaker. Note that if the decoder board is mounted inside your radio, you can use the original speaker. I used a Motorola speaker, but there's no reason why

| | | |
|---|---|---|
| C1,C2 | 0.1 $\mu$F | 272-1069 |
| C3 | 4.7 $\mu$F, 15 volts | 272-1001 |
| C4 | 2.2 $\mu$F | 272-1040 |
| C5 | 100 $\mu$F, 15 volts | 272-1005 |
| C6,C7,C8 | 0.01 $\mu$F | 272-131 |
| Q1 | 2N718, 2N2222, 2N3565 | 276-2009 |
| Q2 | small scr | 276-1059 |
| U1 | NE567V | 276-1721 |
| CR1 | 1N4735, 6.2-volt, 1-watt | 276-561 |
| CR2 | Red LED | 276-041 |
| RY1 | *miniature magnetic reed relay* | 275-230 |
| | *or sensitive mechanical relay* | 275-004 |
| SW1 | dpdt miniature toggle switch | 275-1546 |
| R1 | 10k trimmer | 271-218 |
| R3 | 5k trimmer | 271-217 |

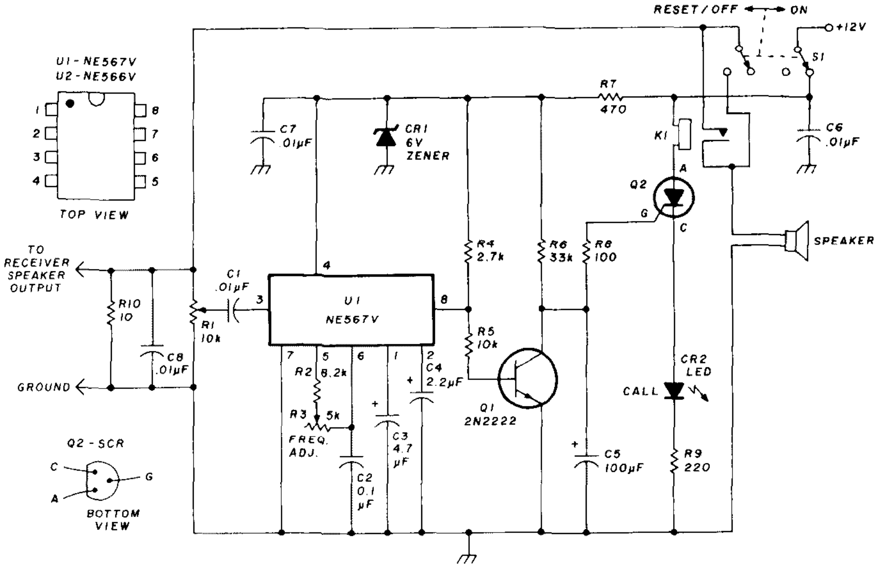


**fig. 1. *Priva-Call* decoder schematic. Capacitors less than 1 $\mu$F are disk ceramics; those over 1 $\mu$F are electrolytics.**

**table 1.** Possible tone frequencies available for use with the *Priva-Call* system using the *Touch-Tone* pad.

| pushbutton numbers | frequency (Hz) |
|---|---|
| 1 and 2 | 697 |
| 4 and 5 | 770 |
| 7 and 8 | 852 |
| * and 0 | 941 |
| 1 and 4 | 1209 |
| 2 and 5 | 1336 |
| 3 and 6 | 1447 |

you couldn't mount a small speaker inside the same minibox as the decoder. See **fig. 4**.

The system I used incorporated both the *Touch-Tone* pad and the decoder board. See **fig. 5**.

## adjustment

You can either use a signal generator set to your tone frequency, or a repeater-group member or friend who will transmit the tone.

**Signal generator method.** Disconnect the *Priva-Call* from the speaker hjack of your radio and connect it to your signal generator. Set the generator for about 1 volt output at the tone frequency you want. Adjust R1 for a reading of 0.1 to 0.2 volt at pin 3 of U1. Now slowly adjust R3 until the system triggers on. The voltage at pin 8 of U1 should decrease to near zero.

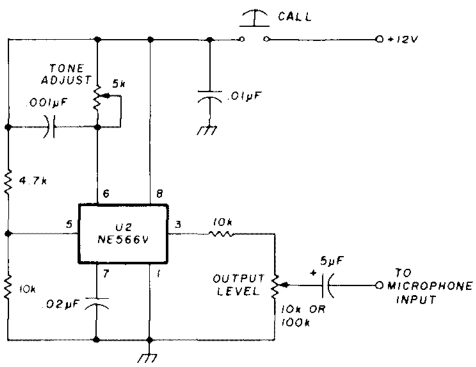


**fig. 3. Encoder system 3. This system uses the NE566V phase-locked loop. Output level is adjusted so that the tone is the same amplitude or slightly less than that of your voice as someone listens to your signal.**

Remove the tone signal and the SCR should remain conducting and the CALL light should remain illuminated. Move switch S1 to RESET then to ON, and the CALL light should turn off, indicating standby mode. If the system triggers falsely when connected to your radio, make sure that the voltage at pin 3 of U1 is not more than 0.2 volt. This is ac voltage, by the way, and *not* dc voltage.

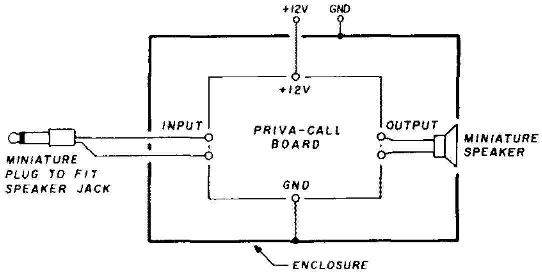


**fig. 4. Self-contained version of the *Priva-Call*.**

**Group-member method.** Instruct your friend to transmit the tone for three seconds then wait for seven seconds repeatedly until proper operation is obtained. Adjust the volume control on your radio for normal listening level. Then adjust R1 and R3 as

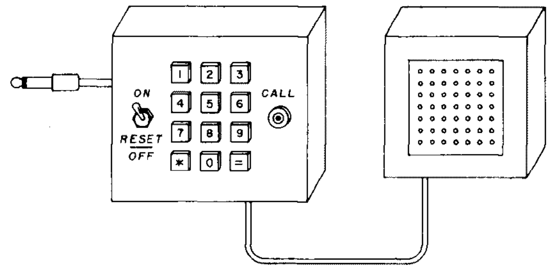


**fig. 5. *Priva-Call* system with external speaker, which is built into a *Touch-Tone*® box (see text).**

described above until your speaker is triggered on every time.

## operation

Place toggle switch S1 in the RESET/OFF position and adjust your radio for normal listening volume. To call a station, press the CALL button or the proper *Touch-Tone* buttons. Hold the CALL button in for at least two seconds so you're sure the other party's speaker has latched on . Now make your call as normal.

To place your *Priva-Call* in the standby mode, move the switch to the RESET/OFF position and adjust your radio for normal volume. Turn on the squelch and move the switch to the ON position. The *Priva-Call* is now ready to receive the proper call tone.

## reference


1. Request a copy of the *LM-555 Timer Application Notes* from Signetics, 811 East Arques Avenue, Sunnyvale, California 94086. See also *The Linear Data Book*, National Semiconductor Corporation, 2900 Semiconductor Drive, Santa Clara, California 95051 (price: $3.50). Also available through Radio Shack stores.


**ham radio**

# measuring resistance values below 1 ohm

## A simple, effective circuit for low-value resistance measurements using readily available parts

**While researching the Amateur Radio** literature I found suprisingly few articles on measuring low values of resistance, especially for measurements under 1 ohm. When making a shunt for a typical 0-15 milliammeter so that it will read 0-150 milliamperes (or higher), we usually need a shunt or resistance with a value of less than 1 ohm and probably close to 0.3 ohm. Unless such a shunt is accurately made it's easy to obtain an error of 20 or 30% in the expanded-scale readings of the shunted meter.

A search revealed some significant facts about resistance measurements in general. The Wheatstone bridge, used to measure resistance, measures medium and high resistances but isn't suitable for measuring resistances of less than 1 ohm. Resistances ranging *from a few to* a few-hundred ohms may be measured by using a constant current of the same value flowing through a known and unknown resistance, then measuring the voltage drop across the unknown resistance.

A method for measuring very low resistance values was found in which large values of current (about 50 amperes) were applied to a low-resistance circuit, then measurements were made of the low value of voltage across the unknown low resistance. Very low values of voltage, divided by high values of current, will give values of resistance in thousandths-of-an-ohm range. This latter method seemed like a good approach, but I decided to use values of current much, much, lower than 50 amperes to measure voltage in the millivolt range.

For these measurements I used a millivoltmeter that reads 0-50 mV on one scale and 0-500 mV on the other, a Simpson model 260 multimeter (to read 100 mA, and a couple of 1-ohm 1% resistors.)

While the next statement may be old stuff to many, it nevertheless bears repeating. The ohm may be defined as the resistance of a conductor through which an emf of 1 volt will maintain a steady current of 1 amperes.* We'll use this principle in a circuit in which the voltage-to-current ratios will be 1 ohm, but the actual values of voltage and current used will be one-tenth of 1 volt and one-tenth of 1 ampere. We will have the following relationships:

$$R = \frac{E}{I}\ ;\ 1\ ohm = \frac{0.1\ volt}{0.1\ ampere}$$

$$\begin{matrix}unknown\ R \\ (less\ than\ 1\ ohm)\end{matrix} \leqslant \frac{millivolt\ reading\ (volts)}{0.1\ ampere}$$

### measurement circuit

The circuit used to make these measurements is shown in **fig. 1**. If you make measurements below 1 ohm, you may wish to use eight binding posts and a minibox enclosure (mine measures 4 by 4 by 2 inches, or 102 by 51mm). Use good-quality binding posts; *do not* use Radio Shack no. 274-661, which are made in two pieces and break apart when the hex nut is tightened. A 1-inch (25mm) spacing on centers for the unknown binding posts will allow resistance measurements in terms of ohms per inch or ohms per mm. This circuit is simple. However, the mechanical connections must be tight, joints well soldered, and no. 12 or 14 (2.1 or 1.6mm) copper wire used to make connections.

### procedure

In making resistance measurements, a 1-ohm, 1% resistor (the "standard") is connected in the unknown position and the battery and milliammeter are connected to their posts (**fig. 1**). The 50-ohm pot is adjusted so that milliammeter reads 100 mA. The millivolt meter is connected last, and it will read 100 mV.

Disconnect one lead of the millivolt meter from its post, remove the 1-ohm, 1% "standard" resistor, and connect the unknown resistor (assumed to be less than one ohm). Gingerly touch the lead of the millivolt meter to its post to see if the millivoltmeter will read on the scale. Observe the millivolt reading

*An interesting but slightly different definition is given in reference 1. **Editor**

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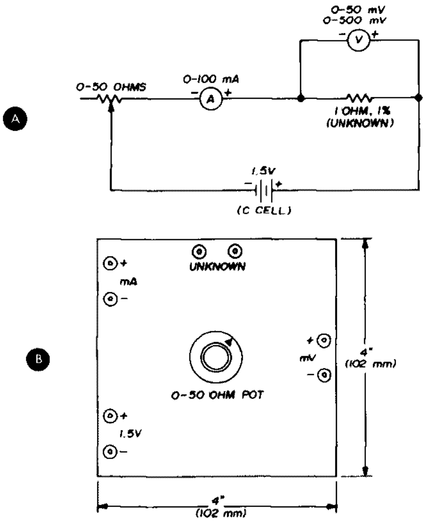


**fig. 1. Schematic, A, and panel arrangement, B, of the low-resistance measuring circuit. Binding posts for the unknown resistance can be spaced 1 inch or 25mm, which allows resistance measurements in terms of unit length. Short lead lengths of large-diameter wire are necessary for accuracy. Insulate binding posts from the chassis.**

and, at the same time, see that the milliammeter reads 100 mA.

The millivolt reading divided by 0.1 is the resistance of the unknown being measured. Remember that if you use clip leads or long wire leads from the unknown resistance to the binding posts their resistance contributes to the value of resistance being measured.

A word of caution. If you are using a sensitive and/or expensive millivoltmeter, be sure the scale or meter setting will accommodate the reading. For example, if you use a 0-25 or 0-50 millivolt range meter and the unknown resistance turns out to be 0.8, 0.9 or 1 ohm, the meter pointer will slam against its stop and meter damage will probably result. Always use the largest millivolt scale first.

Keep in mind also that accuracy is limited by such factors as the accuracy of the 1-ohm, 1% "standard" resistor and the accurracy of the meters and their shunts. Since most good-quality current- and voltage-measuring equipment available to amateurs runs about 1 or 2%, the low-resistance measurement values should also be in that range.

## reference

1. E. A. Mechtly, "The International System of Units, Physical Constants and Conversion Factors," NASA Publication SP-7012, Scientific and Technical Information Division, Office of Technology Utilization, National Aeronautics and Space Administration, Washington, D. C., 1960, page 5. (Available from Superintendent of Documents, U.S. Government Printing Office, Washington, D. C. 20402; price 30 cents.)

**ham radio**

# tone-burst generator for repeater accessing

## Unique, single IC tone-burst generator that can fit inside a microphone case

As more and more repeaters are licensed, interference increases when a common input frequency is shared by multiple repeaters. One solution to this problem is tone-burst accessing. This type of accessing usually requires a short burst, about one second in duration, at a specified frequency such as 1800 Hz. Each time the transmitter is energized, a tone is transmitted which is decoded at the proper repeater.

### circuit description

A very simple, inexpensive, and small tone-burst generator has been developed which will fit inside a microphone case and yet is powered from the transceiver without any additional wiring. The tone-burst circuit uses a CMOS quad NOR gate, CD4001AE. These chips are obtainable from parts suppliers listed in the back of this magazine for less than $1.00. The quad NOR gates are connected in a standard astable multivibrator circuit to generate the tone needed for accessing a repeater. The Tone burst on-time is developed by an R-C time constant which is triggered by the PTT switch closure to ground. A large-valued-power supply capacitor is used to power the multivibrator for a time period slightly longer than the tone burst on-time. The capacitor is normally charged to 12 volts through the PTT circuitry during receive.

**Fig. 1** shows the schematic of the complete tone burst generator. The astable frequency is defined from the equation

$$f = \frac{1}{2.2\ R1C1}\ Hz$$

where R1 is in ohms and C1 is in farads. By making R3 greater than 2R1 the tone-burst frequency is essentially independent of temperature effects and power supply voltage. The astable multivibrator is gated on by another section of the quad NOR gate. The on-time is defined by the R2-C2 time constant and is approximately equal to

$$T_{on} = 1.1\ R2\text{-}C2\ seconds$$

where R2 is in ohms and C2 is in farads. C2 is only allowed to charge after the PTT switch closure. When C2 is charging, the oscillator will be turned on until the C2 charging voltage rises above the input threshhold of the NOR gate. Note that voltage is removed from the circuitry when the PTT line is grounded. To provide a supply voltage during the tone burst interval, capacitor C3 is used to store enough charge to power the astable. C3 is a miniature electrolytic capacitor whose value should be greater than 30$\mu$F at 15 Vdc. The series diode prevents the discharge of C3 when the PTT line is grounded.

A simple T-network is used to attenuate the 12-volt oscillator output voltage to a level necessary to drive the microphone input circuitry of the transmitter. The values of the network should be adjusted to provide sufficient drive voltage without unnecessary loading of the microphone element.

### interfacing

Fig. 2 shows the schematic of the tone-burst generator connected to the microphone and PTT circuitry. Note that only three wires are needed to interconnect the tone-burst generator with the transmitter. The power supply voltage is taken through the PTT relay in the transmitter.

When adjusting the tone-burst frequency, it is helpful to ground pin 1 or 2 of the CD4001 so a tone

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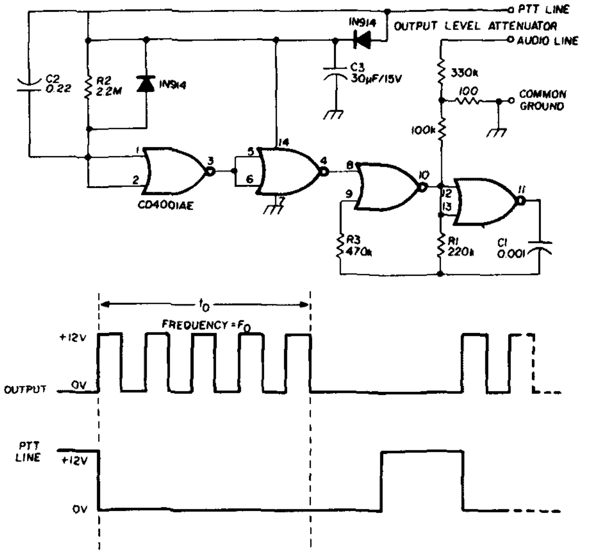



fig. 1. Schematic diagram of the tone-burst generator. A burst is generated each time the microphone switch is closed. Capacitor C3 is used to power the generator.

is generated at all times. The value of R1 should then be adjusted to put the oscillator right on frequency. A pot could be used during adjustment and then replaced after the correct resistance value is found, or if a miniature trimpot is used, it could be left in the circuit at all times.

Since the circuit uses only one IC and a few components, it will easily fit inside the microphone case

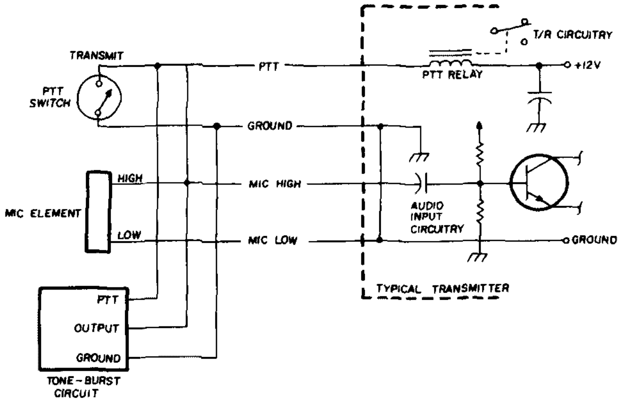
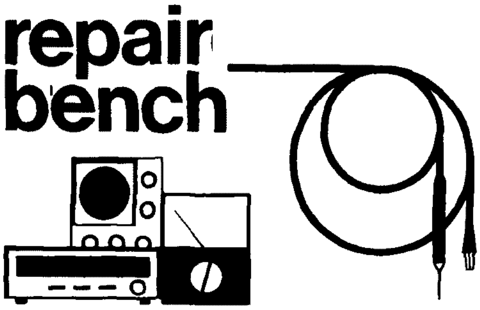



fig. 2. The connections between the tone-burst generator and a transmitter. Power is derived from the PTT circuitry. The generator will easily fit into the transmitter or even the microphone case.

or inside the transmitter chassis at the microphone connector site. If needed, a miniature switch could also be used to disable the circuit when not needed or to change the value of R1 to generate a different frequency.

**ham radio**




# Joe Carr, K4IPV

## troubleshooting the power supply

**It's probably safe to say** that few circuits fail as often, in almost every type of equipment, as the power supply. In fact, a dictum often given younger electronics people by more experienced hands is to look for the power supply problem before attempting to find any others. I know this is a rather large claim, but it's supported by almost two decades of electronic service experience and will probably be corroborated by others who have such experience.

In this month's *repair bench* we'll discuss the troubleshooting techniques appropriate to receivers, transceivers, low-to-medium-power transmitters, and other items of station equipment. It's assumed that the reader is limited as to available test equipment, so the procedures are given in terms of simple voltmeters and ohmmeters, although it's recommended that you also obtain at least a low-cost oscilloscope.

### full-wave supply

Consider the rather ordinary full-wave supply of **fig. 1**. This circuit is of a type frequently encountered in amateur radio equipment. We'll not spend any time describing the operation of this type of circuit because this is an article on troubleshooting, not theory. If you want a review, see the *Radio Amateurs Handbook* and the other appropriate amateur literature.

**By Joseph J. Carr, K4IPV,** 5440 South 8th Road, Arlington, Virginia 22204

The rectified output voltage waveform, without filtering, resembles a series of pulsating dc waves (or inverted parabolas). But with capacitor C3 connected the waveform more nearly resembles that of **fig. 2A**. The output waveform taken across bleeder resistor R1 should be very close to a straight line and be ripple-free.

### common problems

Several different types of problems will be found in power-supply circuits. Those most often found are:

**1.** Hum or ripple on the dc output.

**2.** No output voltage, but the fuse is okay.

**3.** Fuse blows.

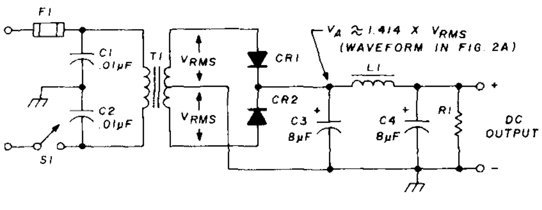



**fig. 1. Typical dc power supply using full-wave rectification. The text describes methods for troubleshooting problems that cause ripple or hum in the output.**

Ripple appears in any number of ways, depending mostly on the nature of the circuit drawing power from the affected power supply. In a communications or broadcast receiver, for example, the hum will most likely be heard in the speaker or earphones; while in a transmitter, the hum will most likely modulate the output and be heard by others. It doesn't only affect ssb/a-m transmitters, incidentally, so also suspect ripple in many types of problems in CW and

fm transmitters. On a television screen, ripple will usually show up as either single (60 Hz) or double (120 Hz) horizontal black bars, sometimes of low contrast, that seem to migrate vertically up the screen.

In almost all cases where ripple or hum originates in the power supply, it's reasonably safe to "pre-indict" the filter capacitors. These components are the most frequent source of trouble. Realize, however, that not all sources of hum or apparent ripple are the fault of the power supply. There are several other sources for such spurious signals.

One of these is a resistive short circuit in a vacuum tube. The heaters, and sometimes the cathode are powered directly from low-voltage ac and this voltage will either appear on the grid or modulate the cathode-plate current directly in some cases. Another case is low-frequency oscillation with a frequency that approximates that of the power line or its second harmonic. Also, in many types of equipment, the possibility of ground loops must be considered as well as common-mode rejection problems or shielding defects as a source of hum.

One differentiating technique is to examine the frequency of the hum. In all equipment that uses any form of *fullwave* rectification, the ripple frequency in the power supply is 120 Hz (or twice the line frequency for those outside the United States and Canada). If the hum is nearer to the line frequency, then you can suspect one of the other causes and temporarily forget the power supply.

These articles on power-supply servicing contain much useful advice for newcomer and old timer alike. Especially appropriate are the author's remarks on "dos and don'ts" in working with this type of circuit. Joe Carr's remarks are *must reading* for anyone thinking about working on power supplies, regardless of the voltages involved. **Editor**

The waveforms in figs. **2A** and **2B** were taken from the same power supply, on the same oscilloscope, with the same sensitivity setting, during the same session. The actual power supply circuit was not unlike that of fig. **1**, but the filter capacitors were higher in value. The waveform of fig. **2A** was taken at point **A** (in fig. **1**) with capacitor C3 connected and in good condition. The waveform in fig. **2B**, on the other hand, represents the waveform at the same point but with C3 disconnected. This exercise effectively simulates an open capacitor as might be found in actual equipment. Note the rather dramatic increase in ripple amplitude.

**table 1. Steps to follow when troubleshooting a power supply when no dc output is obtained.**

| measurement across | approximate value | check for open circuit if no voltage at* |
|---|---|---|
| **1.** T1 primary | 115 Vac | F1, F1 holder; S1; ac power cord |
| **2.** T1 secondary | high-voltage ac | T1 primary; T1 secondary |
| **3.** C3 | high-voltage dc | T1 center tap; center tap grounding; diodes CR1 and CR2 |
| **4.** Across C4 | high-voltage dc | L1 |

*All checks are performed with an ohmmeter and with the ac power plug removed from the wall outlet. Wait a few seconds, then connect an alligator cliplead across the filter capacitors. Connect the negative or grounded end first, then attach the hot end. Leave the cliplead in place for 25 seconds, then remove it. The supply should be safe for both you and the *ohmmeter*. In the last two steps it may be necessary to remove the two diodes before making resistance measurements, as they can give erroneous readings.

Always check the associated wiring and solder connections for each component! Many bad solder connections are difficult to eyeball, so you may want to re-do each one if no other possible cause is found.

## component substitution

If you don't have an oscilloscope, then you'll have to trouble shoot using the component substitution technique. This technique consists of shunting a known good capacitor across each filter capacitor in its turn. It's important to a) use a filter capacitor with the *same* or *higher* ratings and b) scrupulously observe *polarity markings*.

An aluminum electrolytic capacitor might explode if power is applied in reverse of normal polarity. Even if the danger from shrapnel is reduced, the cleanup afterwards and the pure fright involved should be reason enough to take care!

Capacitor bridging must be done with care, not only as mentioned above, but in the actual manner in which you perform the job. You may well be dealing with potentials that can be lethal, so some safety precautions are necessary. In this one respect I advise you to ignore the professional servicer and listen to some expert advice given to beginners:

**1.** Turn the equipment off.

**2.** Using a screwdriver, or preferably an alligator cliplead, ground the filter capacitor for a few seconds.

**3.** Check with a dc voltmeter to make sure all the charge has been *drained off* the capacitor.

**4.** Using either alligator clipleads or solder tacking, connect the known good capacitor across the suspect component.

If the substitute capacitor makes the symptoms disappear, then make the substitution permanent — *first go through the capacitor discharge procedure* given above.

Many people, even (or perhaps especially) those professionally experienced in electronic servicing, are tempted to solder a new replacement directly across the open capacitor which was used originally. This approach to "repair" seems especially convenient if the bad capacitor is a multi-section chassis-mounted type, and an available replacement is of the tubular type . This is an example of popular, but *extremely poor* practice. It's often the case that the old capacitor will short circuit, which will have a spectacularly bad effect on the future life expectancy of your equipment! This is one of those cases where we can show a good reason for demanding good craftmanship on repair jobs. Shunting a bad capacitor with a good one, then, is only a *diagnostic* method and is definitely **NOT** a *repair* technique!

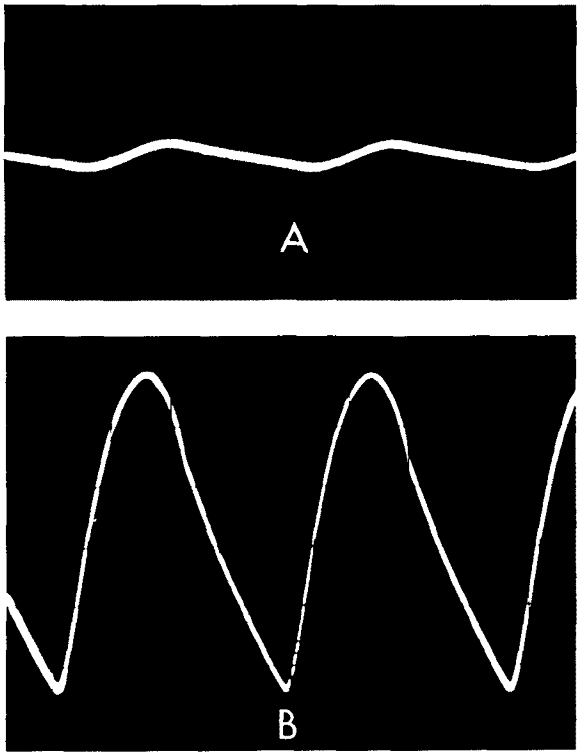



**fig. 2. Oscillograms showing the output waveforms obtained from the supply in fig. 1. Picture A shows the supply output waveform taken with capacitor C3 (fig. 1) connected. Picture B shows the pulsating waves obtained at CR1, CR2 junction (fig. 1), without the succeeding filter, C3, L1, C4 connected.**

In cases where there's no dc output but the fuse is intact, or where the fuse blows, it's necessary to use a vom or vtvm. These instruments are so low in cost these days that you should make one or the other a part of your amateur-station equipment. For most amateur work the vom is preferred. Reasons: portability, low cost, and it's not susceptible to rf fields.

But let's get back to our next problem type: no dc output, but the fuse is intact. **Table 1** shows the steps to follow in this type of job. Set your voltmeter to its highest ac scale and connect the probes across the primary of transformer T1. Turn on the power and observe the voltage reading. If it's too far down the scale to be easily read, reduce the voltmeter range, posittion-by-position, until a good reading is obtained. If you get past the 150-Vac scale, however, and still have no reading, then give up.

A good reading (105-125 Vac in the U.S.) indicates that you should go o n to the next step. If, on the other hand, there's no voltage across T1 primary turn off the set, unplug it from the wall outlet, then use ohmmeter continuity checks (*R x 1* scale) to find the open component. Check those items indicated in the right-hand column of **table 1** for continuity.

Similarly, you must check the voltage across the T1 secondary. Again set your ac voltmeter to its highest ac scale and work down until a readable deflection is obtained. If there's no voltage across T1 secondary, suspect either the secondary or primary winding as being open. Use the same procedure as indicated in the first step.

For the last steps in the procedure, it's necessary to use the dc scales of your voltmeter. The same method of starting with the highest scales and working down is necessary to avoid an expensive surprise.

## the open fuse

A somewhat more spectacular type of power supply defect is the case of the blown fuse. It's reasonably true that "fuses don't cause trouble, they *indicate* trouble." Whenever a fuse blows you must assume that there's a root cause, and that does not usually include a surge on the ac line, as many often believe. To be sure, such activity can occur during a thunderstorm, but in most residential areas few truly destructive surges do occur. Something overloaded that power supply, and it bears at least an attempt at diagnosis!

One of my first employers in electronics was fond of saying that, "All electrical troubleshooting in-

volves the finding of an unwanted path for current, or locating a lost but required path for current," and that "The best tactic is to divide and conquer." Good advice 18 years ago and still good advice today. Such is the approach with the case of the blowing fuse.

We want to find an unwanted path for current (a short circuit) and must follow either (or both) of two schools of thought on how to divide and conquer.

One school maintains that the best method is to obtain a good supply of new fuses and keep replacing them as they blow. In the meantime, between fuse changes, you disconnect first one component then another until one is disconnected that causes the fuses to stop blowing. This technique has a certain validity, but can be expensive, especially if there are a large number of options. A better idea is to mount a television receiver-type ac circuit breaker in a plastic box (well insulated). Solder one end of two alligator clipleads to the breaker. The alligator-clip ends are then used to shunt the blown fuse. This can be a little dangerous if not done correctly, so if you have any doubts, then use up a couple of boxes of fuses.

The other school of thought is to use an ohmmeter to locate the short circuit to ground. This is the method of choice if there is some way of knowing the expected resistance to ground at various critical points. This type of information is often given in amateur equipment service manuals, but in many cases you will have to guess. This technique is the more elegant of the two, provided it works, because one must always be aware that operation of an overloaded circuit, even if the fuse is expected to go up in smoke, can further damage the equipment. Under **NO** circumstances may you use a higher-value fuse!

The actual procedure in "divide-and-conquer" servicing of a power supply circuit depends somewhat upon the type of equipment being serviced. In audio circuits you may have a shorted output tube or transistor, shorted output transformer, and so forth. In a transmitter, on the other hand, a final amplifier operated without protective bias may well blow a fuse if the excitation fails. If an ohmmeter check fails to reveal a short circuit in such a case, then suspect a loss of drive, or a sagging filament, as the cause. The latter problem may occur after a few moments normal operation and it is most frequently seen in transmitters where the final-amplifier tubes are mounted horizontally.

If a short circuit does show up on the ohmmeter, however, it can be conquered by first disconnecting the rectifiers (the most frequent single cause of trouble), and then the load, followed by the filter capacitors. Don't overlook the possibility of a carbon path from an arc or dirt to ground, which could cause the trouble.

## power-supply dos and don'ts

**1.** If at all possible, use an isolation transformer to power all instruments and equipment being used or serviced.

**2.** Always use *well-insulated* alligator clipleads and meter probes. If they are in disrepair then repair them before use.

**3.** *Always* unplug the power cord when connecting or disconnecting clipleads or when soldering tacking.

**4.** Work on a bench with a master power shut-off switch and, if possible, a ground-fault interrupter.

**5.** Use the "buddy-system" and inform the buddy where the master shut-off is located and how it's operated. Similarly, inform all members of your family of the master-switch location and under what circumstances it is to be operated.

**6.** If your equipment and instruments don't have three-wire power cords, install them. In this type of power cord a third wire (usually green in color) is grounded to the equipment chassis.

**7.** *NEVER* work outside or in an area with a concrete or dirt floor* unless the equipment is designed for that purpose by reason of *double-insulation* or *three-wire* power line grounding of the equipment case. Many people have been killed by indoor appliances taken out of doors.

**8.** Now for a seeming paradox: *NEVER* defeat or trust interlocks.

**9.** Never service ac/dc equipment unless operated from an isolation transformer. Similarly, never use ac/dc equipment outdoors.

**10.** Always do quality work, use quality components, and *never* button up a piece of equipment cabinet with a temporary or unorthodox repair inside. A Murphy's Law corrollary states that, "Temporary repairs become permanent if there are more than two screws holding the cabinet together."

**11.** Switch to safety, think safe, work safe, be safe, and live.

*The editor of this article, W6NIF, after 40 years as an amateur, was careless recently when testing a 4500-Vdc power supply. He was working in a garage with a concrete floor and the +4500-volt lead accidentally fell onto the concrete. W6NIF was lucky — he survived.

**ham radio**

# remote checking of repeater shack temperature

By **Fred Johnson, ZL2AMJ,** 15 Field Street, Upper Hutt, New Zealand

**Our repeater site** is situated in a hut at 2823 feet (860m) on top of a mountain. The site is sometimes snow covered in winter, and user stations have often discussed the likely temperature at the site. It was finally decided that a very simple addition to the repeater would enable users to know if the temperature was above a preset level. It would then be possible to track the temperature through the day to learn the time when the temperature passes across that level in the morning, and when it passes back across that level in the late afternoon. Variations in the daily times of change enable estimates of the temperature maxima and minima to be made.

The modification to the repeater is simple. A room thermostat and a capacitor have been added to the repeater, increasing the length of the squelch tail when the thermostat contacts are closed. After the temperature drops below the thermostat setting the tail goes to its longer length condition. In our case, the tail lengthens by 2 seconds when the temperature goes below 40 degrees Fahrenheit (4 degrees Celsius). Our thermostat has been changed several times as the knowledge of the situation improved. The site is rarely visited so the setting we decide upon is permanent for long periods. It is a simple matter to measure the tail length and conclude the present temperature zone. No equipment except a watch is required at measuring stations.

The schematic (**fig. 1**) shows the control circuitry between the receiver mute and the keyed transmitter high-voltage line. Q1 and Q2 are connected as a darlington pair to minimize loading on the CA3089E mute circuitry. Q3 and Q4 are a Schmitt trigger producing a delay time before the transmitter drops out. Q5 is the series transistor which turns the transmitter on and off. The tail length is set by the values of C1, R1, R2 and the input characteristics of Q3. With the addition of C2, (by the thermostat) two very distinct tail length conditions are generated.

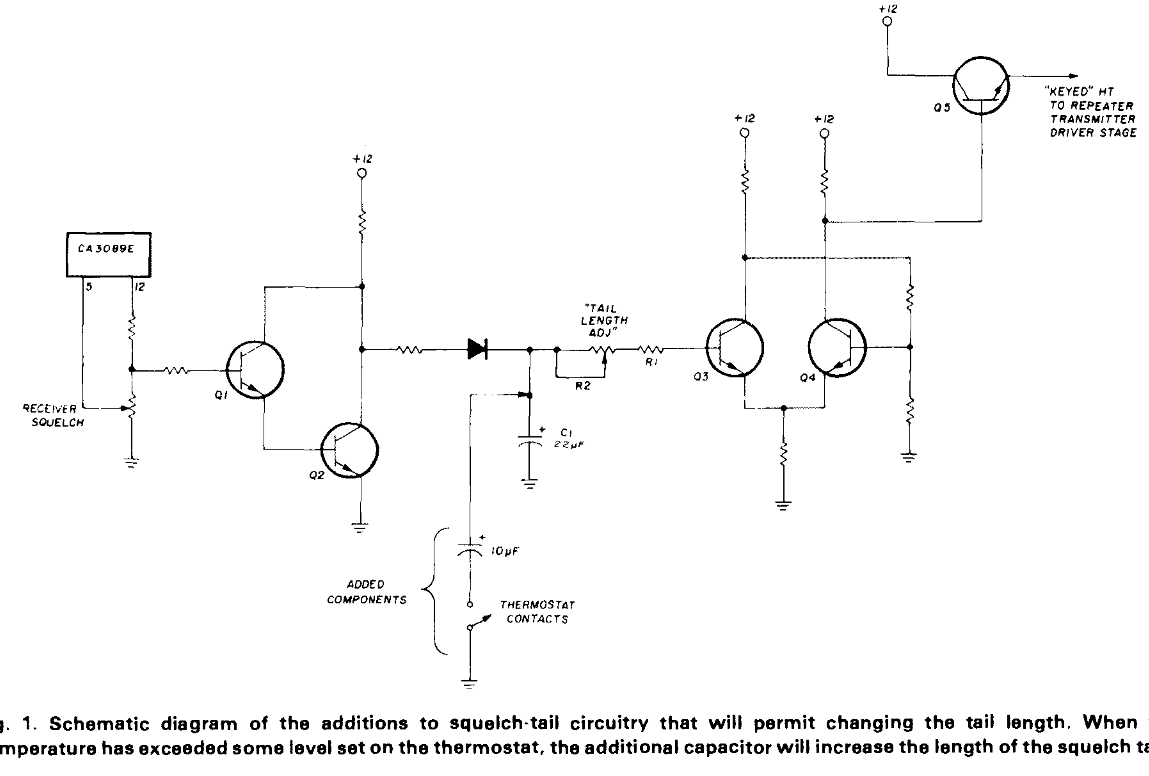



**fig. 1. Schematic diagram of the additions to squelch-tail circuitry that will permit changing the tail length. When the temperature has exceeded some level set on the thermostat, the additional capacitor will increase the length of the squelch tail.**

Although it is unlikely that your repeater will use the same control circuitry as described here, you will probably find some similar points where the addition of an appropriate capacitor and thermostat will cause an increase in the repeater tail length.

**ham radio**

MICROPROCESSORS

# microcomputer interfacing: the 8080 logical instructions

Last month we discussed the concept of, and important use for multi-bit logical instructions such as AND, OR, Exclusive-OR, and COMPLEMENT. This month we'll summarize twenty-eight logical instructions in the 8080A instruction set. It is very important to note that in the case of each logical instruction, *the result is stored in the accumulator*. The previous contents of the accumulator are one of the logical variables in the two-variable logical operation, or in the case of the complement instruction, the only logical variable.

The eight different logical AND instructions, each with the mnemonic ANA S, have the following general form:

| | | |
|---|---|---|
| 1 0 | 1 0 0 | s s s |
| Arithmetic and logical class of instructions | AND operation | 3-bit binary code for source register |

The three bits designated by sss correspond to the register or contents of a memory location that logically operate on the accumulator contents,

| register | octal code | 3-bit register code |
|---|---|---|
| B | 0 | 000 |
| C | 1 | 001 |
| D | 2 | 010 |
| E | 3 | 011 |
| H | 4 | 100 |
| L | 5 | 101 |
| M | 6 | 110 |
| A | 7 | 111 |

**By David G. Larsen, WB4HYJ, Peter R. Rony, and Jonathan A. Titus**

Mr. Larsen, Department of Chemistry, and Dr. Rony, Department of Chemical Engineering, are with the Virginia Polytechnic Institute and State University, Blacksburg, Virginia. Mr. Jonathan Titus is President of Tychon, Inc., Blacksburg, Virginia.

The OR and Exclusive-OR instructions, which have the mnemonics ORA S and XRA S, respectively, have the same general form as the ANA S instruction byte. Thus, for the XRA S instruction the instruction byte is,

| | | |
|---|---|---|
| 1 0 | 1 0 1 | s s s |
| Arithmetic and logical class of instructions | Exclusive-OR operation | 3-bit binary code for source register |

and for the ORA S instruction,

| | | |
|---|---|---|
| 1 0 | 1 1 0 | s s s |
| Arithmetic and logical class of instructions | OR operation | 3-bit binary code for source register |

Some examples are:

| logical operation | mnemonic | octal instruction code |
|---|---|---|
| B•A→A | ANA B | 240 |
| M•A→A | ANA M | 246 |
| A•A→A | ANA A | 247 |
| C+A→A | XRA C | 251 |
| L+A→A | XRA L | 255 |
| A+A→A | XRA A | 257 |
| D+A→A | ORA D | 262 |
| E+A→A | ORA E | 263 |
| M+A→A | ORA M | 266 |
| A+A→A | ORA A | 267 |

Another logical instruction, the complement accumulator instruction, has the mnemonic CMA A and the octal instruction byte 057.

In preceding columns,[1,2] we discussed the con-

cept of an *immediate instruction*, a multi-byte instruction that contains the desired data within the instruction. The three immediate logical operations can be summarized in the following way:

| logical operation | mnemonic | octal instruction code |
|---|---|---|
| <B2>·A→A | <ANI B2> | <346 B2> |
| <B2>+A→A | <XRI B2> | <356 B2> |
| <B2>+A→A | <ORI B2> | <366 B2> |

In the two tables, the symbol → means "is replaced by". Thus, the notation B·A→A means that we AND the variable B with the variable A, and then replace the original contents of A by the result of the logical operation. Within the 8080A microprocessor chip, the logical operation is performed in a temporary accumulator, with the logical result in the temporary accumulator being *copied* into the accumulator register, A.

In last month's column we demonstrated one use for logical instructions, the testing of flag or comparator bits associated with the on/off state of external devices. The AND multi-bit operation is particularly useful when it is desired to clear, filter, or *mask* specific bits in an input data byte. For example, consider the ASCII code for the numeric characters 0 through 9:

| character | octal ASCII code | binary ASCII code |
|---|---|---|
| 0 | 260 | 10 110 000 |
| 1 | 261 | 10 110 001 |
| 2 | 262 | 10 110 010 |
| 3 | 263 | 10 110 011 |
| 4 | 264 | 10 110 100 |
| 5 | 265 | 10 110 101 |
| 6 | 266 | 10 110 110 |
| 7 | 267 | 10 110 111 |
| 8 | 270 | 10 111 000 |
| 9 | 271 | 10 111 001 |

Once the ASCII code is in the microcomputer, the most significant four bits are of little use and can be *stripped* away from the data byte. A simple program that accomplishes such a task is:

| LO memory address | octal instruction code | mnemonic | comments |
|---|---|---|---|
| 000 | 333 | IN | Input ASCII numbers from the following device |
| 001 | 015 | 015 | Device 015 |
| 002 | 346 | ANI | AND the accumulator contents with the following data byte |
| 003 | 017 | 017 | Mask byte that *masks* the most significant four bits in the ASCII word |

The program accomplishes the following Boolean operation for ASCII 5:

| ASCII 5 | | Mask byte | | BCD data of interest |
|---|---|---|---|---|
| 10110101 | • | 00001111 | = | 00000101 |

This logical result of the AND operation is, 00000101. This form contains a single BCD digit per input data byte, with the BCD digit being the least significant four bits in the byte, D3-D0. The remaining four bits, D7-D4, can be used to store another BCD digit provided there's some means to position this second digit in these open bit positions. If the four bits of storage space are not taken advantage of, 50 per cent of the memory capacity will be washed.

To *pack* two BCD digits into a single data byte, you must have the capability to rotate the contents of your accumulator. As an example, the rotate left instruction, which has the mnemonic RLC, the octal instruction byte 007, and can be described as follows: "The content of the accumulator is rotated left one position. The low order bit and the carry flag are both set to the value shifted out of the high order bit position".[3] The four rotate instructions in the 8080 instruction set have been previously discussed.[4] The accumulator is the only register that can be rotated in an 8080A chip. Other registers are rotated simply by moving them to the accumulator register, performing the necessary rotation operations and then returning the rotated byte back to the original register. Besides shifting BCD digits back and forth in data bytes, important uses for the rotate instructions will appear when discussing decision-making operations.

A simple program that can be used to pack two BCD digits into a single data byte is listed below.

| LO memory address | octal instruction code | mnemonic | comments |
|---|---|---|---|
| 000 | 333 | IN | Input ASCII 5 from the following device |
| 001 | 015 | 015 | Device 015 |
| 002 | 346 | ANI | Mask off the four most significant bits |
| 003 | 017 | 017 | Mask byte |
| 004 | 007 | RLC | Rotate the BCD digit |
| 005 | 007 | RLC | into the four most |
| 006 | 007 | RLC | significant bits that have |
| 007 | 007 | RLC | just been cleared |
| 010 | 107 | MOV B,A | Store this result in register B |
| 011 | 333 | IN | Input next ASCII character, in this case ASCII 7, from the following device |

| | | | |
|---|---|---|---|
| 012 | 015 | 015 | Device 015 |
| 013 | 346 | ANI | Mask off the four most significant bits |
| 014 | 017 | 017 | Mask byte |
| 015 | 260 | ORA B | OR contents of register B with contents of accumulator |
| 016 | 167 | MOV M,A | Store packed data into memory, the location being specified by the contents of the H,L register pair |

The result of this sequence of steps is the data byte, 01010111, stored in memory. The four most significant bits are BCD 5, and the four least significant bits are BCD 7. Observe the use of the ORA B instruction, which permitted the combination of two data bytes into one, without changing either. Special 8080 microcomputer programs, called simulators, are available that permit you to follow the execution of an 8080 program step by step by observing the changes in the contents of the internal registers.* If applied to the above program, you would observe the following, *after the execution of the indicated instruction bytes*:

*One such program, called *DEBUG*, has been developed by Tychon, Inc., in Blacksburg, Virginia; it requires the use of a teleprinter or CRT.

| executed instruction bytes | accumulator | register B |
|---|---|---|
| IN 015 | 10110101 | — — — |
| ANI 017 | 00000101 | — — — |
| RLC, RLC, RLC, RLC | 01010000 | — — — |
| MOV B,A | 01010000 | 01010000 |
| IN 015 | 10110111 | 01010000 |
| ANI 107 | 00000111 | 01010000 |
| ORA B | 01010111 | 01010000 |

This completes our discussion of the more important logical instructions in the 8080A instruction set. Additional examples will be used in the following columns, where they will be incorporated into data-manipulation and decision-making tasks.

## references


1. Jonathan A. Titus, David G. Larsen, and Peter R. Rony, "Microcomputer Interfacing: The MOV and MVI 8080 instructions," *ham radio*, March, 1977, page 74.
2. David G. Larsen, Peter R. Rony, and Jonathan A. Titus, "Microcomputer Interfacing: Register pair instructions," *ham radio*, June, 1977, page 76.
3. Intel Corporation, *Intel 8080 Microcomputer Systems User's Manual*, Intel Corporation, Santa Clara, California, 1975.
4. Paul E. Field, David G. Larsen, Peter R. Rony, and Jonathan A. Titus, "Microcomputer Interfacing: A Software UART," *ham radio*, November, 1976, page 60.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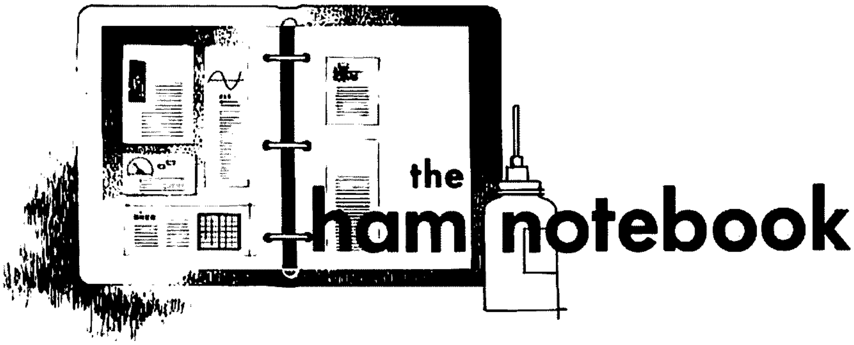
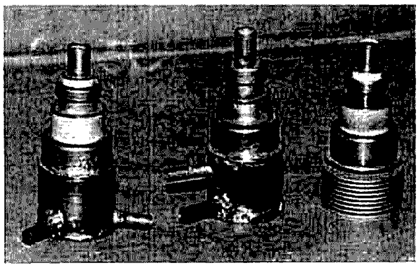



## water-cooled 2C39

Why water cool a 2C39 tube? The answer is to prolong its life. This is written with the intent of using 2C39s in a Motorola T44 converted for ATV. T44s were not made for long periods of transmitting, as in TV but with use, watercooling the 2C39 tubes never even get warm. Also, with water-cooled tubes, 900 volts on the plate of the amplifier will give you a healthy output and yet the tube will not be working hard. The 6146 or 2E26 tubes (whichever model you have) should have some air because the box they are in will get very hot without it.

I have put quite a few of these tubes together and the best method I have found is to use a water tank 5/8 or 1 inch (16 or 25mm) long and 1-1/8 inch (28.6mm) in diameter. I use a piece of 1-1/8 inch (28.6mm) O.D. copper tubing cut to the length I want. Then I cut a piece of copper flashing (any thin piece of copper sheet can be used) to cover one end of the tubing. The short tank is for the later model T44 which uses a vane for tuning, while the larger tank is for the early T44 using a plunger on top of the tube for tuning. Note: the vane type must have the inlet and outlet on top and the plunger type has the inlet and outlet on the side.

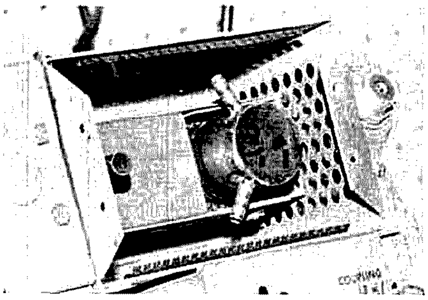
For the inlet and outlet, I use a short piece of 3/16 inch (4.75mm) tubing about 1/4-inch to 3/8-inch (6.35 to 9.5mm) long, just long enough to put a plastic hose on. The brass tubing can be found in most hobby shops that carry model airplanes and supplies. The plastic tubing is the same as used for tropical fish tanks. After the inlet and outlet tubes have been soldered in place, the next step is to take the heat sink off the 2C39. There are two types of heat sink. One has an Allen screw while the other has a right hand thread. On the type with the thread, I hold the heat sink in a vise and hold the plate of the tube with a pair of pliers (not too tight) and gently turn it off. With the heat sink off, all that remains to finish the job is to solder the tank to the plate connection of the tube. I use a 250-watt electric solder bolt and soft solder, being very careful not to get the tube too hot. After a little soldering, I put it under running water to cool, then I solder some more and let the tube cool until the job is done. Don't solder for "looks." Solder for a tight job without too much heat on the tube. I do not use acid core solder, just regular rosin core solder.



**Three 2C39 tubes include an unmodified tube (right) with original finned heat sink in place. Tube on left is from early T44 (plunger tuning) and tube in center shows "dry run" or first try at modification. Note slightly different placement of water inlet and outlet tubes.**

Now don't get carried away as I did. I had the T44 tuned, so I made both tubes, put them in and filled them with water. I was so far out of tune by making all of the changes that, no matter what I tried, I couldn't get any output; so it was back to the drawing board for me.

The way that seems to work the best here is to peak the unit up at the frequency I plan to use, then take output readings in two positions, write them down and use them for reference. I use the original power supply for this. Then I pull out the 2C39 tripler, modify it, reinstall it and retune the input and output stages to get the same or better output. On the later T44, it may take a little adjusting of the Z2 and Z4 shorting bars, but keep in mind not to let the tube get too hot. At this point in the procedure there is still no water in the tube. When the output is again satisfactory treat the 2C39 amplifier the same, but keep in mind not to test a long time as you will ruin the tubes. I also mounted one tank a little off-center, putting it somewhat closer to one side of the cavity. The result was that I increased the output from 10 watts to 15 watts. This does differ from the



**Water-cooled 2C39 in place T44 final amplifier compartment. Plastic tubing for admitting and carrying away coolant not yet attached to short pieces of brass tubing on water jacket.**

tube-to-tube, so you may not get exactly the same readings.

Now the tubes are ready for the water but, again, take the tripler first and retune it, as the water will make a big difference (use only distilled water). After the tripler stage is in tune, run the water through both the tubes and retune every stage from the 6146 (or 2E26) through the final. Once the tubes are filled with water and the circuits tuned, you should have no more detuning troubles.

I use a "Little Giant" lawn ornament pump in a plastic three-gallon container. Half an hour of continuous running doesn't even begin to warm the water. We now run our amplifier with 900 volts on the plates. Air cooling the 6146 (or 2E26) is a very big help, and removing the cover also helps without causing any output loss.

My thanks to K9CZI for the photo work.

**Daniel J. Smies, WA9RPB**

## inter-band calibration stability for the Collins R-388 (51-J)

The Collins R-388 is a marvelously accurate piece of gear. It is possible to read frequency to an accuracy of about 400 Hz between 0.5 and 30.5 MHz. The heart of the receiver is a very linear and very stable permeability tuned oscillator. If you follow the technical manual for calibrating band-by-band — and there are thirty of them — you can maintain this accuracy. There can develop, however, a small problem connected with calibrating the receiver. This problem does not affect accuracy if proper procedure is followed each time the band is changed, but it slows calibration process and detracts from the receiver's tidiness of operation.

Calibration of an even band holds very well for all the even bands, but is off by a few kHz on each of the odd bands. For example, as you go from band-to-band you hear almost perfect zero-beating on even bands and tones of approximately the same pitch, *e.g.*, 2 kHz on alternate bands. If you don't use the bfo to read precise frequency, there is no problem for you can read frequency quite closely just by using the MHz and kHz dials and tuning by ear. Except for the vfo, all the tuned circuits in the receiver are in the same position before and after you follow the procedure to make odd and even bands calibrate (index) at the same point.

There are four frequency-determining elements in the calibration circuits: The calibration crystal, the crystals associated with the first mixer (selected one at a time by the band-selector switch), the vfo, and the bfo. I checked each of these to see if any pulling occurred when I rotated the band selector switch band to band. There was none. I therefore ruled out circuit design as a cause of the calibration differences between odd and even bands.

Instead, I found the problem to be a matter of adjustment not covered in the manual. The problem was located in the relationship between the vfo and the bfo, the only *adjustable* frequency-determining elements.

On even bands, the oscillator crystal frequency is subtracted from the vfo frequency; and on odd bands, the vfo frequency is subtracted from the oscillator crystal frequency. Therein lies both the problem and its solution. Below is a chart which shows what happens, for example, on band 7 (6.5 to 7.5 MHz) and band 8 (7.5 to 8.5 MHz).

### Calibration on Band 8

| | | |
|---|---|---|
| 10 | MHz | (output from local crystal oscillator) |
| -8 | MHz | (harmonic of 100 kHz calibration crystal) |
| 2 | MHz | |

Therefore, the vfo mid-scale frequency *should* be 2.5 MHz in order to produce the 0.5 MHz final i-f, but let's assume that it is *actually* 2.49 MHz. Then

| | | |
|---|---|---|
| 2.49 | MHz | (vfo frequency) |
| -2.00 | MHz | (difference frequency: oscillator crystal frequency minus calibration crystal frequency |
| 0.49 | MHz | to final i-f |
| -0.49 | MHz | (bfo frequency to produce zero beat and indexing at 8.00 MHz). |

Now, without touching either bfo or vfo, change to band 7 and note:

### Calibration on Band 7

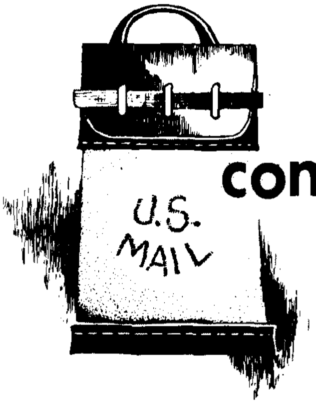
| | | |
|---|---|---|
| 10 | MHz | (output from local crystal oscillator) |
| -7 | MHz | (harmonic of 100 kHz calibration oscillator) |
| 3 | MHz | |

### Subtracting vfo frequency

| | | |
|---|---|---|
| 3.00 | MHz | (difference frequency: local crystal frequency minus calibration crystal frequency) |
| -2.49 | MHz | (vfo frequency) |
| 0.51 | MHz | (to final i-f) |
| 0.49 | MHz | (bfo frequency unchanged from band 8) |
| 0.02 | MHz | (beat frequency audible on band 7) |

A very simple solution exists. Place a pickup loop near the calibration oscillator tube and connect it to another receiver capable of tuning the frequency range between 2-3 MHz. A small multiband radio can be used with good results. Tune for a harmonic of the calibration crystal. Switch on the bfo and listen for a beat note. Rock the bfo knob slightly, and make sure you are hearing the bfo, not some spurious signal. Line up the bfo knob index with the indexing mark on the cabinet. Adjust the bfo coil for zero beat. This will set the bfo to 0.5 MHz and cure the difficulty.

**George Hirshfield, W5OZF**




# comments

## microwave amplifier design

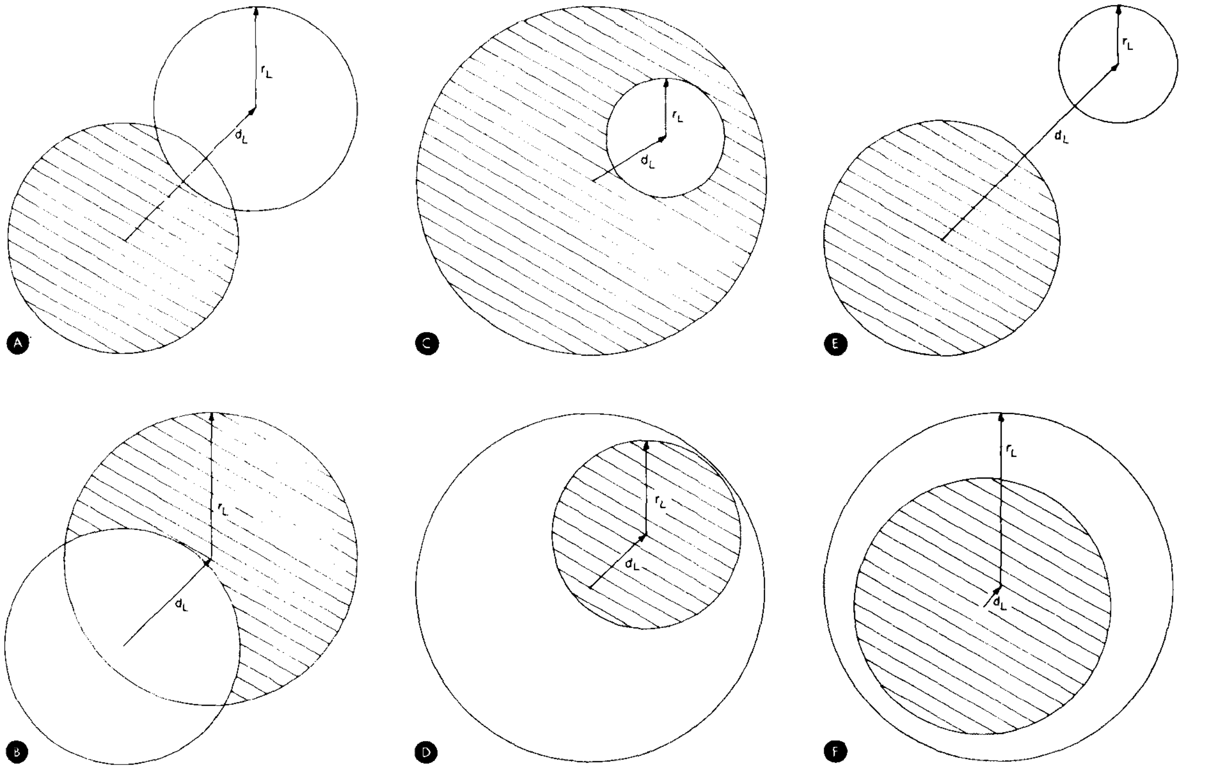
**Dear HR:**

I feel compelled to correct what I consider to be a serious error in Paul Shuch's otherwise well written article on "Solid State Microwave Amplifier Design" in the October, 1976, issue of *ham radio.*

Under the heading "Gain and Stability Analysis," Paul states that "If $K$ (Rollet's stability factor) is greater than 1, the amplifier will be stable under any combination of input and output impedances or phase angles." This statement is incorrect, although it is understood how it is easy to make such a sweeping statement from a reading of HP Application Note AN-154 (Paul's reference 5) alone.

This fundamental error could be the reason why many amplifiers exist today which are only marginally stable, depending on antenna or load connections, despite their designer's belief that the amplifier is "unconditionally stable." The crux of the matter is that $K$ greater than unity is a *necessary* condition for unconditional system stability, but not a *sufficient* one. Stability analysis of uhf amplifiers is far from as simple as Paul suggests.

First, it must be noted that the expression for the *device* stability factor, as I prefer to call it, is *independent* of either source or load impedance. To ensure a stable design, it




CONDITIONALLY STABLE $K < 1$ — CONDITIONALLY STABLE $K > 1$ — UNCONDITIONALLY STABLE $K > 1$

**fig. 1. Stability circle analysis contributed by VK3TK.**

is necessary to know the frequency range over which the system is potentially stable and the load and source impedances which can be used to give stable operation over this frequency range. This information requires that the device stability factor $K$ be known over the frequency range of interest and the reflection coefficients $S_{11}$ and $S_{22}$ for the terminated network (these are not the device s-parameters).*

$$S_{11} = S_{11} + \frac{S_{12} \cdot S_{21} \Gamma_L}{1 - s_{22}\ \Gamma_L} = \frac{S_{11} - \Delta\Gamma_L}{1 - s_{22}\ \Gamma_L}$$

$$S_{22} = s_{22} + \frac{s_{12} \cdot s_{21}\ \Gamma_S}{1 - s_{11}\ \Gamma_S} = \frac{s_{22} - \Delta\Gamma_S}{1 - s_{11}\Gamma_S}$$

for $\Delta$ = determinant of device scattering matrix, *i.e.* $s_{11} \bullet s_{22} - s_{12} \bullet s_{21}$.

Since the source and load terminations being considered are passive networks, their reflection coefficients $\Gamma_S$ and $\Gamma_L$ will be less than unity. For a two-port network to be unconditionally stable, it is necessary that $|S_{11}| < 1$ for all $\Gamma_L$ as $\Gamma_L$ is changed arbitrarily, but kept so that $|\Gamma_L| < 1$. Similarly, it is necessary that $|s_{22}| < 1$ for all $\Gamma_s$ as $\Gamma_s$ is changed arbitrarily, with $|\Gamma_S| < 1$.

Consideration of the $S_{11}$ and $S_{22}$ equations shows that if $|s_{11}| > 1$, then any $\Gamma_L$ will cause $|s_{11}| > 1$ and the network is potentially unstable for all $\Gamma_L$ and the given $\Gamma_S$. Stability with respect to the input port will only then be obtained by ensuring that the positive real part of $Z_s$ is greater than the negative real part of the input immitance. For the condition $|s_{11}| < 1$, the magnitude of $s_{11}$ is less than unity for any passive $\Gamma_L$. Further consideration of the two equations shows that the whole $\Gamma_L$ plane can be separated into two regions, one for which the input immitance is positive real — the stable region, and the other for which the input immitance is negative real — the unstable region.

*A capital S is used to denote external network S-parameters; a lower-case s is used to describe device parameters.

The boundary between these two regions can be defined by solving the relationship

$$|s_{11}| = 1.$$

Using

$$|S_{11}|^2 = S_{11} \cdot s_{11}{}^* = \frac{s_{11} - \Delta\Gamma_L}{1 - s_{22}\Gamma_L} \cdot$$

$$\frac{s_{11}{}^* - \Delta^*\Gamma_L{}^*}{1 - s_{22}{}^*\Gamma_L{}^*} = 1$$

it can be shown, with some algebraic difficulty, that the stable and unstable regions of operation are defined by a circle in the $\Gamma_L$ plane (unit circle) where:

$$\textit{center } d_L = \frac{C_2{}^*}{|s_{22}|^2 - |\Delta|^2}$$

$$\textit{radius } r_L = \frac{|S_{21}S_{12}|}{|S_{22}|^2 - |\Delta|^2}$$

where $d_L$ is located on a line through $S_{22}{}^*$ and the origin of the unit circle, and $C_2$ is as previously defined in WA6UAM's article.

Typical examples of stability circles are shown in the diagrams to the left. The region of the $\Gamma_L$ plane which provides a positive real input impedance (*i.e.* $|s_{11}| < 1$) is indicated as follows:

**1.** If the stability circle *includes* the origin of the unit circle, the *inside* of the stability circle (within the unit circle) defines the area in which a selected $\Gamma_L$ will result in a positive real input immitance.

**2.** If the stability circle *excludes* the origin, then the area of the unit circle *outside* the stability circle is the area of positive real input immitance.

The stability of the output port can be investigated wiith respect to $\Gamma_S$ plane being given by:

$$\textit{center } d_s = \frac{C_1{}^*}{|s_{11}|^2 - |\Delta|^2}$$

$$\textit{radius } r_s = \frac{|S_{12} \cdot S_{21}|}{|S_{11}|^2 - |\Delta|^2}$$

The necessary conditions for a two port to be absolutely stable can now be stated: A two-port network is absolutely stable if there exists no passive source or load termination which will cause the system to oscillate. This is equivalent to requiring the unstable regions to lie outside the unit circles in the $\Gamma_S$ and the $\Gamma_L$ planes. This is satisfied if

$$|d_s| - |r_s| > 1$$
$$|d_L| - |r_L| > 1$$
$$|s_{11}| < 1\ |s_{22}| > 1.$$

The establishment of the possible regions of unstable operation inside the unit circle is a necessary prelude to the application of any design technique. Without knowing the constraints imposed on the system by stability requirements, it is pointless to proceed to determine the source and load reflection coefficients to meet some particular gain specification.

I hope this very brief resume has helped in some way to clear the air on this subject.

**Graham J. Clements, VK3TK**
**Technical Director**
**Relcom Engineering**
**Melbourne, Australia**

*Graham Clement's letter is as fine an exposition on stability-circle analysis as I've read since William Froehner's article in the October 16, 1967 issue of* Electronics. *And Mr. Clements goes further than that article by correctly pointing out that K > 1 is a necessary, but not a sufficient condition, for absolute stability.*

*Although the stability circle analysis approach outlined by Mr. Clements appears entirely correct, I regard it as frosting on the cake. It is my belief, and confirmed by others, that the only conditions for absolute stability are $K > 1$, $S_{11} < 1$, and $S_{22} < 1$. In other words, an amplifier with $K > 1$ can oscillate* only *at the* design frequency *if either the input or output impedance is negative. Since the rest of my design equations fall apart if $S_{11}$ or $S_2$ are* greater than 1, there is little danger of inadvertently designing an oscillator using the formula in my article.

*The key here, of course, is the term "at the design frequency." Any transistor having $S_{11} < 1$, $S_{22} < 1$, and $K > 1$ at a design frequency may well*

*exhibit K < 1, $S_{11}$ >1, or $S_{22}$ > 1 at some far removed frequency. Thus an amplifier which is unconditionally stable over a particular passband may indeed oscillate at some other frequency! This is another reason to use interstage isolators as described in the February, 1977, issue of* ham radio *(page 26), even for "unconditionally stable" amplifiers.*

*Although I have not performed a rigorous analysis to prove that the three conditions for absolute stability are always K > 1, $S_{11}$ < 1, and $S_{22}$ < 1, it has been proven empirically in countless amplifier designs by myself and others. I would be very interested in any careful analysis of this question which* ham radio *readers may care to undertake.*

**H. Paul Shuch, WA6UAM**
**San Jose, California**

## antenna noise bridge

**Dear HR:**

The article on the improved RX noise bridge in the February, 1977, issue of *ham radio* was very well done; authors Hubbs and Doting have come up with an excellent solution to the accuracy problem of the original design by YA1GJM (*ham radio*, January, 1973). When designing and building antennas, RX measurements are a must and, considering the simplicity and accuracy of this improved noise bridge, my advice is, "Don't leave home without it!"

The range-extender idea is a very nice way to get added coverage for this instrument, especially for 80- and 160-meter work. For those using 300- or 600-ohm line the thought occurred to me that another version of the range extender assembly might be made except in this case the resistor would be placed in parallel with the *unknown* impedance instead of in series. For best accuracy the resistor should be nearly equal to the resistance of the pot, say 220 to 240 ohms, and the assembly should be constructed using as physically small a resistor as possible to keep down added stray capacitance.

One word of caution: (especially to hand-held calculator wielders) don't impute any greater accuracy to the computations than that of your original readings. If your reading accuracy was good to within 5%, the computed result isn't going to be any better just because you have it out to eight decimal places. This comment applies to either range-extension computation.

**Forrest E. Gehrke K2BT**
**Mountain Lakes, New Jersey**

*Mr. Gehrke's suggestion for using a 220-ohm shunt range extender with the RX noise bridge is an excellent idea. The 100-ohm series resistor is*

**measured impedance of adapters**

| series adapter (output shorted) frequency (MHz) | $R_p$ | $C_p$ | shunt adapter (output open) frequency (MHz) | $R_p$ | $C_p$ |
|---|---|---|---|---|---|
| 3.5 | 101 | 0 | 3.5 | 165 | 5 |
| 7.0 | 100 | 0 | 7.0 | 165 | 5 |
| 14.0 | 100 | 0 | 14.0 | 165 | 4 |
| 21.0 | 100 | −1 | 21.0 | 165 | 3 |
| 28.0 | 100 | −2 | 28.0 | 165 | 3 |

**Note: The small $C_p$ offsets shown above are used to correct the $C_p$ readings to have a 220-ohm resistor available. I used a 170-ohm resistor in the shunt adapter.**

| frequency (MHz) | $R_p$ | $C_p$ | shunt adapter? | series adapter? | measured series impedance $R_s$ | measured series impedance $X_s$ | actual impedance $R_s$ | actual impedance $X_s$ |
|---|---|---|---|---|---|---|---|---|
| **test load 1 (350-pF capacitor)** | | | | | | | | |
| 3.5 | 203 | 107 | yes | yes | 2 | −131 | 0 | −130 |
| 7.0 | 140 | 102 | no | yes | ~1 | −63 | 0 | −65 |
| 14.0 | 109 | 33 | no | yes | ~1 | −31 | 0 | −32 |
| 21.0 | 103 | 15 | no | yes | ~2 | −21 | 0 | −22 |
| 28.0 | 102 | 6 | no | yes | 0 | −14 | 0 | −16 |
| **test load 2 (14-pF capacitor)** | | | | | | | | |
| 3.5 | 165 | 20 | yes | no | 0 | −3000 | 0 | −3200 |
| 7.0 | 165 | 20 | yes | no | 0 | −1500 | 0 | −1800 |
| 14.0 | 165 | 19 | yes | no | 0 | −750 | 0 | −800 |
| 21.0 | 165 | 18 | yes | no | 0 | −500 | 0 | −540 |
| 28.0 | 165 | 18 | yes | no | 0 | −380 | 0 | −400 |
| **test load 3 (11.6 feet of RG-58/U, open circuited)** | | | | | | | | |
| 3.5 | 237 | 230 | no | yes | −3 | −116 | 4 | −121 |
| 3.5 | 163 | 396 | yes | no | 1 | −115 | 4 | −121 |
| 7.0 | 152 | 433 | yes | no | 1 | −53 | 2 | −50 |
| 7.0 | 122 | 86 | no | yes | 1 | −47 | 2 | −60 |
| 14.0 | 101 | 0 | no | yes | 1 | 0 | 1 | 0 |
| 21.0 | 130 | −27 | no | yes | 8 | −48 | 3 | 50 |
| 21.0 | 150 | −125 | yes | no | 2 | −58 | 3 | 50 |
| 28.0 | 148 | 3 | yes | no | ~1400 | 0 | ~1500 | 0 |
| **test load 4 (1000-ohm carbon resistor)** | | | | | | | | |
| 3.5 | 142 | 5 | yes | no | ~1000 | 0 | ~1000 | 0 |
| 7.0 | 142 | 5 | yes | no | ~1000 | 0 | ~1000 | 0 |
| 14.0 | 142 | 4 | yes | no | ~1000 | 0 | ~1000 | 0 |
| 21.0 | 142 | 3 | yes | no | ~1000 | 0 | ~1000 | 0 |
| 28.0 | 142 | 3 | yes | no | ~1000 | 0 | ~1000 | 0 |

useful, as explained in our article, for measuring high Q (low resistance) terminations. We offered no suggestion for high resistance terminations; Mr. Gehrke's solution fills this void quite nicely. With a shunt extension device, it's possible to bring these high resistance terminations within the range of the bridge. In fact, using either and sometimes both the series extender and/or the shunt extender, it is theoretically possible to measure any impedance at 3.5 MHz and higher. Frank and I have built a shunt range extension assembly to prove the suggestion is practical. Our findings summarized below support that conclusion.

1. The shunt range extender can be made physically using the same PL-295 connector and the same SO-239 Motorola pin-plug adapter as suggested in our article for the series device. The only difference is that a short length of wire is used to connect the center terminals together, and the resistor is connected from center pin to shield.

2. There is about 5 pF of stray capacitance to ground and about 25-30 nanohenries of series inductances in the finished unit. These strays cause the noise bridge null to shift about 5 pF in the capacitive direction when using the shunt device. This offset can be compensated for with sufficient accuracy (in most cases) by merely subtracting the offset from the readings one obtains.

It is interesting to note that these same strays exist in the series range extender. However, nature conspires to make them functionally transparent in this case. The input impedance of the series extender with the output short-circuited is very nearly a pure resistance. This is caused by the fact that 25 nanohenries in series with 100 ohms resistance is functionally equivalent to the same 100-ohm resistor in parallel with a negative capacitor. This negative capacitor nicely compensates for the stray capacitance in the circuit. The same compensation effect does not exist for the shunt assembly.

3. Besides being a nice theoretical technique, the shunt extender works in practice as the following data shows.

We feel Mr. Gehrke's suggestion is a valuable addition to noise bridge and impedance measuring technology for the ham. Our findings demonstrate the idea is also practical for implementation by the amateur.

**Bob Hubbs, W6BXI**

## wideband preamp

**Dear HR:**

The Article in the October issue of *ham radio* on the "Wideband Preamp" by W1AAZ was intriguing and yet simple enough for me and two friends to quickly build three models. Unfortunately, the article didn't give us enough indication of the preamp's performance and I thought some readers might be interested in our results.

A Motorola HEP S3013 was used in place of the 2N5109 while the balun was wound using 10 twisted, bifilar turns of no. 30 AWG (0.25mm) wire on a 5//16-inch (8mm) Q-Z core. The balun seemed to be the most critical part of the design. My first attempt with 8 turns of no. 26 AWG (0.4mm) resulted in considerably less gain than the final results. The circuit was laid out on a 1-1/4 x 1-3/4-inch (3.2x4.5cm) printed circuit board. A 22 V battery was used to power the preamp.

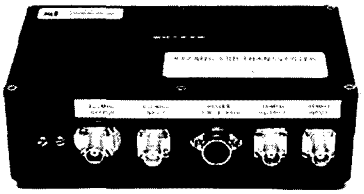
The results were quite surprising. With a 50-ohm signal generator on the input and a 50-ohm termination on the output, there was a minimum of 10 dB voltage gain over the range of 2 to 70 MHz. The noise figure was measured to be less than 3dB from 1.5 to 30 MHz. Although no measurements were recorded, a quick check of desensitization and intermodulation distortion showed very good results. My thanks to W1AAZ and *ham radio* for bringing this design to my attention.

**Glenn S. Williams WB2DHG**
**Oakhurst, New Jersey**

# NEW products

For literature on any of the new products, use our *Check-Off* service on page 142.

## 144 and 432 MHz linear transverters



Microwave Modules, Ltd., of Liverpool, England, has introduced a line of transverters (transmitting converters) that are of great interest to Amateurs who want to operate on the vhf bands. Although the term transverter is usually applied to a circuit that only transmits, these units have a receiving converter built in as well. Thus, only a hf-band transceiver need be connected to the Microwave Modules box to enable you to operate on the higher bands with ease.

Three units of primary interest to Amateurs are the MMT 144/28, MMT 432/28, and the MMT 432/28 Mark 4. The first number in the designation indicates the frequency of the band of operation, in this case either 144 MHz or 432 MHz. The second number indicates the frequency of the input (or output) signal required for mixing (or as an i-f output). Thus the 432/28 will allow operation in the Amateur 432-434 MHz range, with an input of 28 MHz for transmitting and an i-f output of 28 MHz for receiving.

The MMT 432/28 Mark 4 is of special interest since it has been broadbanded to cover a 4-MHz range. This feature has been incorporated to allow you to operate both weak-signal (432 MHz) and the future Oscar 8 (436 MHz). Two additional units have also been recently introduced, the MMT 432/50 and the MMT 438/ATV. Power output from the transverters is nominally 10 watts, PEP. Input and output impedance is 50 ohms, with BNC fittings on the enclosure for connection. There is a separate connector for the 28-MHz i-f output to the receiver, marked 28-MHz OUTPUT. A connector is provided for a separate 144-MHz (or 432-MHz, as the case may be) input, but the connector is not wired up. Instructions are given to enable the user to connect this input jack if desired. Normally, the 432- or 144-MHz input jack serves as both transmitting and receiving connections. Separating the two functions would be useful if you wanted to drive a linear amplifier while transmitting. PIN diodes perform the internal switching function.

DC power required for the transverters is in the range of 12 to 14 volts for both units. Quiescent current for the 144-MHz model is 300 mA; for the 432 unit it is 180 mA. Current drain rises to approximately 2 amperes on peaks for both units.

Drive power required for full output on both models is 500 milliwatts, but there is an internal attenuator that may be jumpered out of the circuit to allow the use of an input as low as 5 milliwatts of drive.

Other features worth noting are a receive converter noise figure of approximately 2.5 dB (144 MHz) and 3 dB (432 MHz); a cast-aluminum enclosure for good shielding and mechanical stability; and a crystal oscillator that starts high enough to avoid the need for a large number of multipliers and their spurious products — 101 MHz for the 432 unit and 116 MHz for the 144 transverter. Both boxes measure 7-3/8 inches wide, 2-1/4 inches high, and 5-1/2 inches deep, including connectors (18.7x5.7x13.9cm). Suggested list prices for the transverters start at $199.95 for the MMT 144/28, $229.95 for the MMT 432/28, and $249.95 for the MMT 432/28 Mark 4. The source for this equipment is Spectrum International, Incorporated, Post Office Box 1084, Concord, Massachusetts 01742.

## full-feature frequency counter

Here's a high accuracy frequency counter for those working within the Citizen Band and Amateur disciplines. The counter has recently been made available from Communications Power, Inc. Designated model CPI FC-70, the frequency counter features a bright seven-digit LED readout with anti-glare louvers — great when you're working in a dimly lit environment.

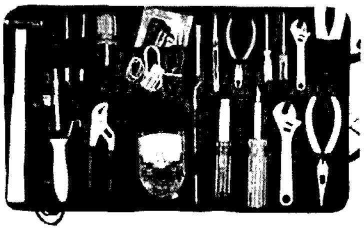
Resolution is within 10 hertz; accuracy is rated at 0.0003 per cent, which is considerably higher than the FCC's 0.005 per cent requirement. The FC-70 accepts 400 watts of throughput power. It has a high-

impedance input, which means it's easily used with rf oscillators and grid dippers. It's also useful for testing i-fs, filter characteristics, and crystal response.

The FC-70 operates from either 12 Vdc or 115 Vac. Quick disconnect cables are supplied for both voltages. The FC-70 has a guaranteed upper frequency limit of 40 MHz; 55 MHz is typical. Looks like a nice piece of test equipment for the serious technician working with high-frequency communications equipment.

For more information on the CPI FC-70 counter, as well as information on CPI's complete product line, write Mr. Robert Artigo, Communications Power, Inc., 2407 Charleston Road, Mountain View, California 94043.

## electronics tools in a roll-pouch kit



This new product, offered by Jensen Tools and Alloys, looks like the answer to the tool-kit problem for field engineers and electronics technicians. It's called the JTK-81 — a tool kit that contains more than 25 essential tools in a roll pouch that's easy to store in drawer or pocket.

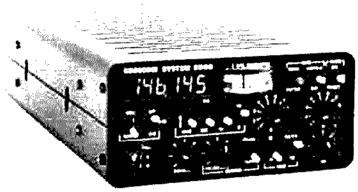
The tool complement consists of pliers, cutters, screwdrivers, nutdrivers, wire strippers, hex and spline keys, soldering equipment, hammer, and more. A Triplett model 310 vom is offered as an optional accessory. The tool package fits neatly into a multipocketed 12 by 21 inch (305 by 533mm) vinyl roll pouch.

The JTK-81 kit without vom is priced at $75.00. With meter, the kit price is $127.00. Quantity prices are significantly lower.

If you'd like a free catalog describing more than 3000 hard-to-find

tools, as well as other Jensen products, drop a note to Jensen Tools and Alloys, 4117 North 44 Street, Phoenix, Arizona 85018.

## two-meter transceiver

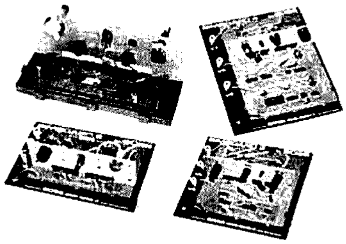


System 3000, a microcomputer-based two-meter transceiver offered by Edgecom, Inc., provides amateurs with a complete personal communications ystem. It has an on-board computer that provides unusual flexibility. Some of its many features:

- Ten front-panel programmable priority channels
- Priority-channel silent monitor
- Built-in scanner
- Two-frequency subaudible tone encoder/decoder
- Transmitter frequency offset
- Audio alarm

For more information on the System 3000, write Edgecom, Inc., 2909 Oregon Court A3, Torrance, California 90503.

## dip breadboard kits



You can obtain DIP Breadboard Kits in three larger models from Hammond Manufacturing Company. The three new models are Bimboard 2, 3, and 4. They consist of individual Bimboards slotted together and mounted onto a 1/16-inch (1.5mm) thick matte-black aluminum base.

The new Bimboard models provide



More Details? CHECK—OFF Page 142

24, 36, and 48 in.$^2$ (155, 232, and 310 cm$^2$) breadboarding area. Included are 1100, 1650 and 2200 individual sockets.

Aluminum backplates, which are mounted on four nonslip rubber feet, are fitted with four screw terminals. Input power and ground leads may be connected to these terminals. Also included are 2-, 3-, or 4-component support brackets, which provide mounting for larger components. For more information write Hammond Manufacturing Company, Inc., 385 Nagel Drive, Buffalo, New York 14225.

## counter-generator with prescaler

Lunar Electronics presents a new frequency counter-generator with a five-digit display and a seven-digit readout with front-panel scaling. It's the model DX-555P — a basic 30-MHz counter with prescaler. The instrument has a 10-MHz time base, which includes easy zero adjust to WWV. The built-in prescaler extends the count range to 300 MHz (activated by a rear-panel switch).

Featured is a variable-frequency marker oscillator, which covers 440 kHz-30 MHz in three bands. When the marker oscillator is activated (front-panel switch), its output is available from a rear-panel jack and is also displayed on the counter readout.

Marker-oscillator output, which may be amplitude modulated, is of sufficient amplitude for aligning receivers with 455-kHz i-fs up through 30 MHz. The high harmonic output may also be used throughout the lower vhf range with careful attention to frequency, which will preclude aligning your receiver on images.

The Model DX-555P with prescaler lists at $239.95. Without prescaler the price is $189.95. For further information write Lunar Electronics, P.O. Box 82183, San Diego, California 92138.



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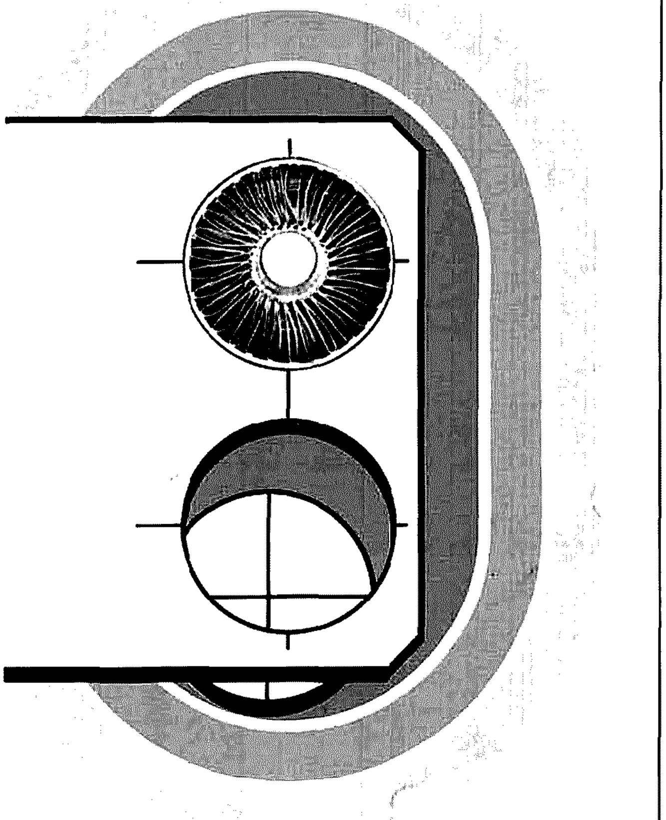
$1.50

# ham radio magazine

**OCTOBER 1977**





## stripline kilowatt for two meters

# ham radio magazine

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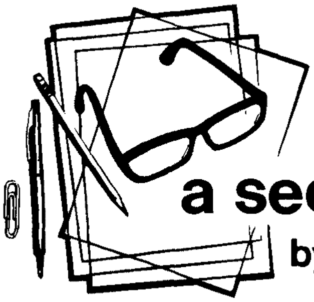


# a second look

by Jim Fisk

**In the last few months there has been** an increase in the number of inquiries and complaints received by the FCC concerning state and local laws which deal with radio and television. Most of the calls and letters are related to CB, and ask whether a particular ordinance is constitutional, or complain about the enforcement of a state statute that is thought to be in conflict with the FCC regulations. To answer these questions, the FCC recently released Public Notice 87276 which provides information to amateurs, CBers, and other FCC licensees who feel they have run into improper local regulation of radio communications.

As the Public Notice points out, as early as 1912 the Congress recognized that radio communications was inherently interstate in nature, was a form of commerce, and was uniquely adaptable to uniform regulation by the federal government. The Radio Act of 1927 increased the federal government's authority to regulate this area, and the passage of the Communication Act of 1934 substantially completed the trend toward comprehensive federal regulation of interstate and foreign communications by wire and radio. To execute and enforce its provisions, the Act also provided for the establishment of the FCC.

It was the primary goal of the Communications Act to make available, "to all people of the United States a rapid, efficient, nation-wide, and world-wide wire and radio communication service . . ." The Act proposed to execute this policy by "centralizing authority" and by "granting additional authority with respect to interstate and foreign commerce in radio communication . . ." Furthermore, the Act claims complete jurisdiction over radio energy for the federal government, as stated in Section 301:

> It is the purpose of this Act, among other things, to maintain the control of the United States over all the channels of interstate and foreign radio transmission; and to provide for the use of such channels . . . under licenses granted by Federal authority . . . No person shall use or operate any apparatus for the transmission of energy or communications or signals by radio . . . except under and in accordance with this Act and with a license in that behalf granted under the vision of this act, 47 U.S.C. §301.

In addition, Congress granted the Commission the authority to establish a pervasive system of regulation in the various radio services. Section 303 of the Act gives such numerous powers to the FCC as to leave no doubt as to the extent of this regulatory scheme. These and other sections of the Act indicate the clear intention of Congress that radio be regulated by the federal government.

Under the Supremacy Clause of the United States Constitution, state and local statutes may be pre-empted when (1) a local law conflicts with a law enacted by Congress, or (2) when Congress has adopted pervasive legislation in a particular field with the intent that regulation in the area will be conducted exclusively by the federal government. Furthermore, local ordinances which unreasonably burden interstate commerce may be invalidated under the authority granted to the federal government by the Commerce Clause of the United States Constitution.

Whether or not a particular local statute has been pre-empted by federal legislation is a question of law, and when a conflict between federal and state law arises, the courts, both state and federal, are usually called upon to make the final decision. For proper resolution the specific local law in question must be reviewed, and each case must be carefully judged on its own facts.

In general it may be said that in matters involving purely local concerns, the courts have found that reasonable local statutes are not in conflict with the Communications Act; such things as local zoning ordinances limiting antenna heights, and the right of local courts to adjudicate property rights involving licensee's facilities have all been upheld. On the other hand, where a local law conflicts with the FCC's regulatory scheme for radio services, the federal law will prevail; state laws involving the censorship of material carried on broadcast stations, for example, and those requiring FCC licensees to refrain from activities required by the Communications Act have been struck down.

The FCC does not have the resources to routinely monitor state and local laws which affect radio communications, nor can the FCC intervene in every local court proceeding in which the validity of various laws is tested. However, amateurs and other FCC licensees who feel victimized by improper local laws may raise federal pre-emption in their own behalf, and local legislative bodies should consider the issue when contemplating the enactment of ordinances in areas regulated by the Communications Act. Since many of the inquiries received by the FCC directly concern CB (and indirectly, amateur radio), local lawmakers should also be aware that the FCC has issued extensive regulations governing this area. Therefore, local ordinances designed specifically to regulate CB (or amateur) transmissions could be invalid according to the legal principles discussed above.

**Jim Fisk, W1HR**
**editor-in-chief**

# presstop

THE TYPE ACCEPTANCE and 10-meter amplifier ban dockets themselves — Dockets 20117 and 20116 — were not discussed at recent FCC meetings, only the ARRL's request for oral argument on those two propositions. After considerable discussion it was decided to reserve the decision on the League's request until later, concurrent with the Commission's actual discussion of the dockets. At this point it seems likely that formal consideration of those two proposals should occur in early October.

MARITIME MOBILES ABOARD U.S. VESSELS gained many new frequencies effective September 12. All U.S. licensed Amateurs operating on the high seas (not in the waters of a foreign power) will be permitted to operate on any frequency that Amateurs are authorized in that ITU Region. For example, in nearby waters (ITU Region 2) maritime mobiles may use all domestic U.S. Amateur frequencies, from 160 up. In the other ITU Regions Amateurs are more limited, of course — a U.S. maritime mobile off the coast of Europe or Africa would be restricted to 7000-7100 on 40 meters, 144-146 on two, and forbidden 6 and 220.

ARMA PRESIDENT DENNIS HAD tendered his resignation to the ARMA membership in a three-page letter. In the letter Denny cited the heavy burden the ARMA presidency had placed on him personally and on Dentron, a burden that — with a few notable exceptions — had not been shared by other ARMA members. Another factor in his resignation decision was criticism from a few of the ARMA members, criticism that reflected on his personal motives and integrity in his efforts to establish and maintain ARMA, while some of these same critical members were the ones whose catering to the illegal CB market had brought ARMA into being in the first place.

Rumors Of A Possible Split in the organization have been in the air recently, with a vocal minority of ARMA members more and more openly critical of the majority's support of the "Amateur equipment for Amateurs" concept. The problem of those members whose "unethical conduct," Dennis said, "has brought on the legal and regulatory problems we now face..." is a prime reason that he decided to resign.

STATE LICENSING OF CBers and Novice and Technician Class Amateurs has been proposed in Michigan by State Senator Basil Brown. Licensing would apply only to radios used in cars, would require a $2 yearly registration fee, a description of the rig along with details of where it was obtained, and a description of the car it is installed in.

The Proposal, Senate Bill 409, exempts General and higher class Amateurs as well as visitors from out of state. It has been referred to the Senate Judicial Committee, which is chaired by Senator Brown. Michigan Amateurs wishing to comment on it can write Senator Brown or their own state senators in care of the Michigan Senate, Lansing, Michigan 48902.

CENTRAL STATES VHF CONFERENCE in Kansas City presented the John Chambers award to W6PO for his outstanding contribution to rf power amplifier design and assistance to VHF and EME oriented people. The F9FT antennas captured honors on both 144 and 432 MHz with gains that measured upwards of 1.0 dB above its nearest competitor which included such notables as KLMs and Quagis. The best 144-MHz noise figure was registered by WAØRDX using a neutralized 2N5297. At 432 MHz, K2UYH's V244 GaAs fet measured 1.25 dB while WB5LUA's entry, an NE645 bipolar, was close behind at 1.35 dB. Next year's plans call for the conference to be held in Rochester, Minnesota starting on August 18.

KH6BZF EXCHANGED 599s with N4DT (WB4PAG) on OSCAR orbit 12634B, culminating a series of skeds which also include K1HTV, K2SWZ, W1JR, WA8UUY, VE3BNO, and W2KLN. Other east coasters wishing a sked with KH6BZF should write including an SASE.

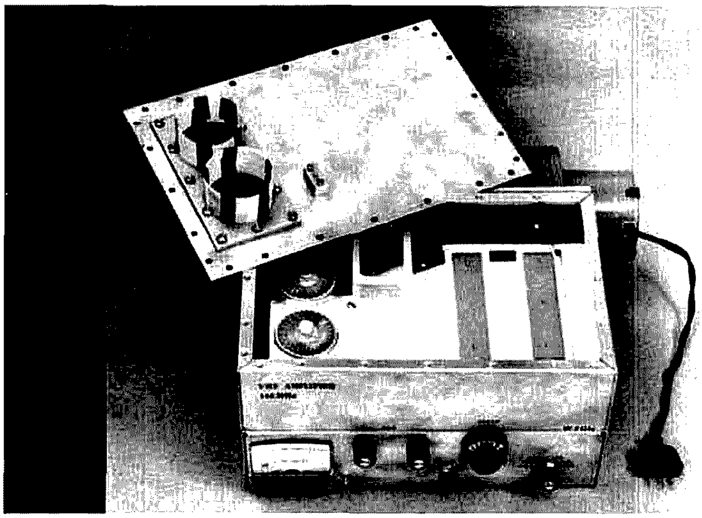
AN ASCII OK and other benefits of the FCC's "bandwidth docket" (20777) may not be far off. Safety and Special Bureau Chief Charley Higginbotham told the audience at a session on regulatory matters at the APCO (Associated Public Service Communications Officers) Convention in Chicago that they could expect a report and order on 20777 to come out "this fall."

INTERFERENCE ON 160 and possibly the high end of 75 meters could result from the FCC's recent approval of wide-band swept anti-theft systems. The three bands authorized for such systems are 1.7-2.3, 4.05-4.95, and 7.4-9.0 MHz, with a maximum field strength of 100 microvolts per meter at 30 meters.

Anti-Theft Systems must not interfere with radio communications, so can be shut down if they bother Amateur operations.

CW SENDING TEST is being dropped by the FCC for all Commission administered Amateur examinations, shortening and simplifying (since examiners won't need CW qualifications) the exam.

Novice Exams Administered by volunteer examiners will still require a sending test, however, to weed out really bad fists.



# stripline kilowatt

## for two meters

Complete layout and construction details for a compact two-meter kilowatt based on the popular K2RIW design for 432 MHz

**The stripline rf power amplifier** designed by K2RIW for 432 MHz and described in *QST* several years ago has gained wide acceptance and use.[1] At the present time it's estimated that 300 to 400 of these amplifiers are in use around the world. The techniques outlined by K2RIW — using a pair of inexpensive ceramic tetrodes in a parallel stripline configuration — were used in a two-meter power amplifier designed by W9OJI.[2] A number of other single-tube stripline power amplifiers for two meters have also appeared in the amateur magazines. In this article I will try to put all this background and experience into construction information for a two-meter power amplifier packaged in the same compact box as K2RIW's original design for 432 MHz. Such a unit could possibly become as popular on 144 MHz as the K2RIW amplifier has on 432.

Several stripline power amplifiers based on the layout described in this article have already been built and thoroughly tested on the air. When operated in the class AB1 linear mode, the amplifier provides 600 watts output with 6 to 8 watts drive. The all-mode two-meter transceivers now on the market have more than ample output to drive this amplifier to full output; the sharply tuned circuits in the amplifier help to attenuate any out-of-band products from the driver.

Any of the tubes from the 4CX250 series are suitable for the amplifier. However, the cooling problem is simplified by using 8930 tubes which are similar to 4CX250Rs except that they have a 2-inch (50mm) diameter anode. The dimensions for the plate line, chimneys, and top cover will be given for both tube types, but 8930s are the recommended choice.

Referring to the schematic, **fig. 1**, the plate line is a quarter wavelength long, with the plate blocking capacitor in the form of a Teflon sandwich at the cold end. Plate tuning is accomplished by a combination of fixed copper plate and beryllium copper flapper capacitor mounted below and near the plate end of the line, **fig. 2**. The capacitive loading to the output is adjusted by a flapper capacitor above the plate line which is also at the plate end of the line.

**By Fred Merry, W2GN,** 35 Highland Drive, East Greenbush, New York 12061

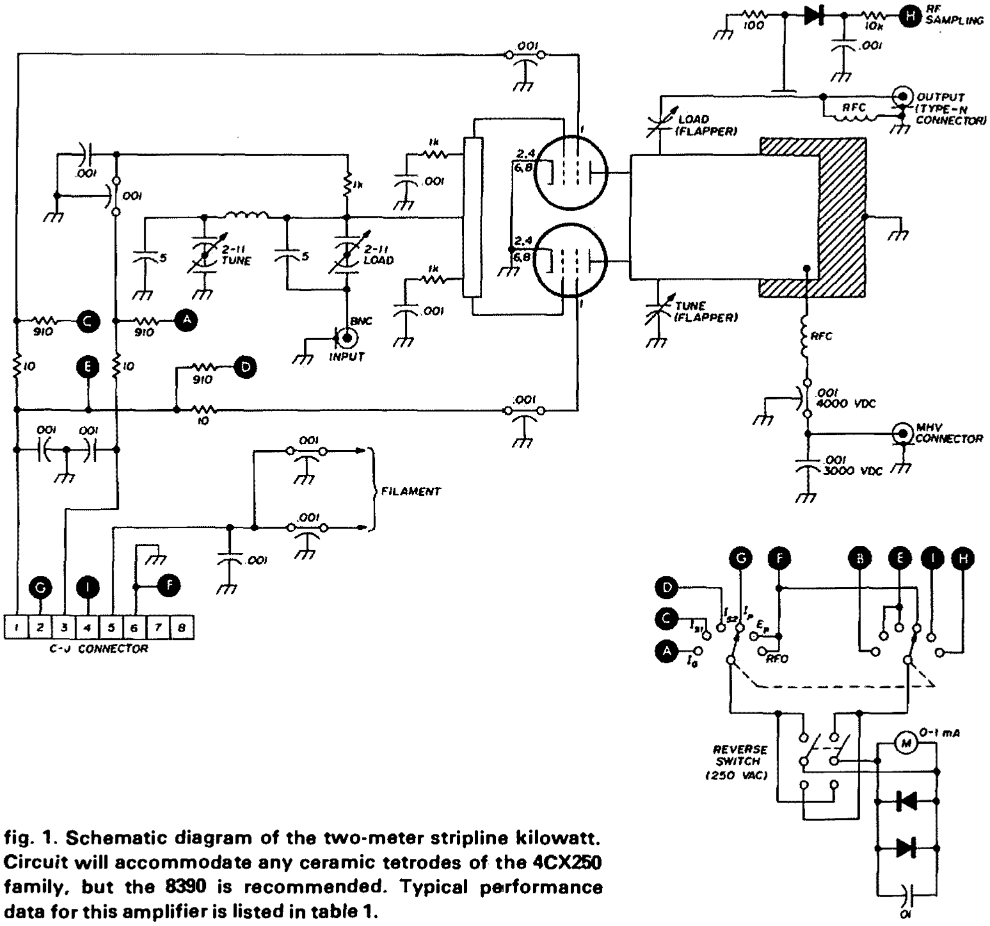
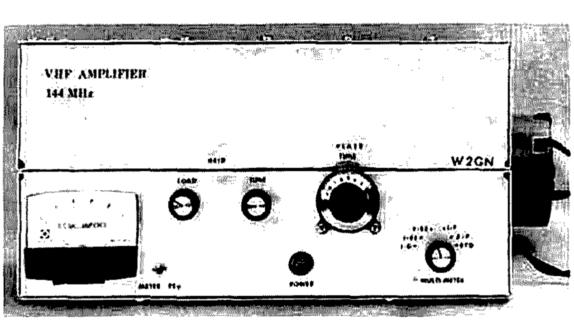



**fig. 1. Schematic diagram of the two-meter stripline kilowatt. Circuit will accommodate any ceramic tetrodes of the 4CX250 family, but the 8390 is recommended. Typical performance data for this amplifier is listed in table 1.**

The grid circuit consists of a 3-turn inductor tuned by a butterfly capacitor at the end away from the grids. Capacitive coupling is used for the input circuit. Eimac 620A or 630A sockets plus the construction and bypassing techniques used provide stability. Stability is further assured by loading the grid circuit down to approximately 300 ohms (derived from the grid bias resistor and two other 1000-ohm resistors mounted close to the grid socket connections of each tube, **fig. 1**).




**Stripline kilowatt for two meters is compact, and measures only 6-inches (15cm) high, 12-inches (30.5cm) wide, and 8-inches (20cm) deep.**

Chassis construction involves the use of three standard chassis: two 8 x 12 x 3 inches (20.3x30.5x7.6cm) and one 5 x 7 x 3 inches (12.7x17.8x7.6cm), plus a top plate and a bottom cover. The chassis preparation is covered by the illustrations accompanying this article which include complete drilling, punching, and cutting details. A good starting point is the top cover which is a piece of 3/32-inch (2.5mm) thick aluminum cut and drilled as shown in **fig. 3**. The large holes are made with a hole saw, 2-1/4 inches (5.7cm) for 8930 tubes, or 1-3/4 inches (4.5cm) for 4CX250s. The vent plate, which is 3/16-inches (4.5mm) thick can also be drilled and hole sawed at the same time. A piece of aluminum screening is cut to the size of the vent plate; the screen is fastened between the top plate and the vent plate with 1/2-inch (12.5mm) long screws, 6-32 (M3.5) lockwashers, and nuts.

The plate loading adjustment block is cut from a piece of 1/2-inch (12.5mm) square aluminum bar stock, drilled and tapped as shown in **fig. 3**. This block is fastened to the top plate with 3/4-inch (19mm) long 4-40 (M3) screws. The 8-32 (M4) nylon adjustment screw is cut to size and inserted in the block. This completes the top cover assembly.

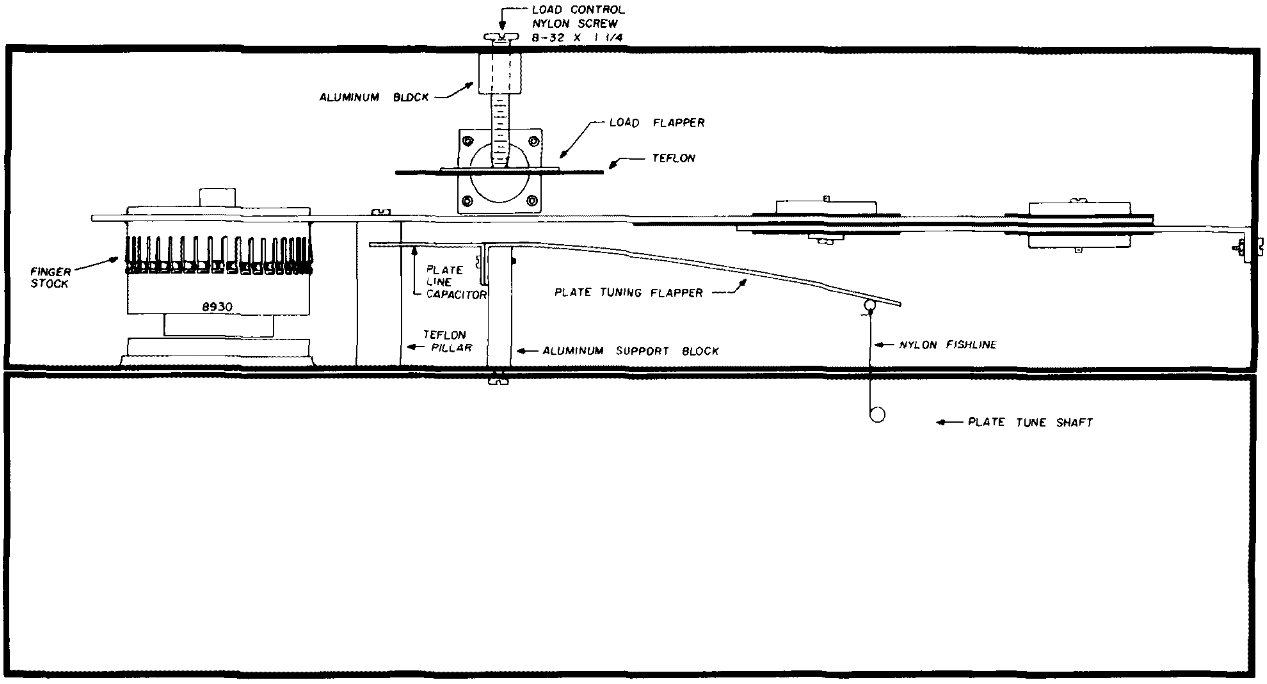
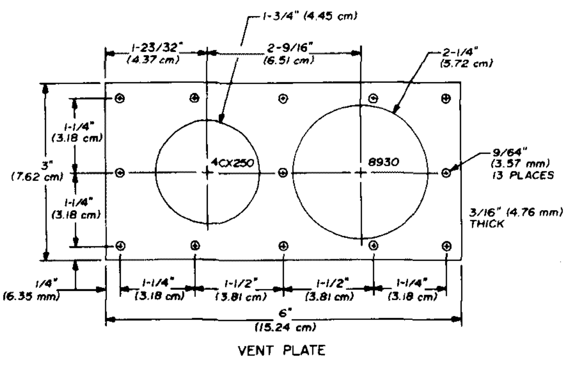



**fig. 2. Cross-sectional view of the two-meter kilowatt showing the location of the plate circuit components.**

The upper chassis is now prepared as shown in **figs. 4, 5,** and **6**. Chassis punches 2-1/4 and 1-3/4 inch (5.7 and 4.5cm) in diameter are required for the socket holes and the air intake. Note also that the plate with a 1-3/4-inch (4.5cm) hole requires the use of a hole saw to get through the 3/16 inch (4.5mm) thickness. This screened air intake plate must be chosen for either a hose-connected blower, or a blower mounted directly on the chassis. The drawing



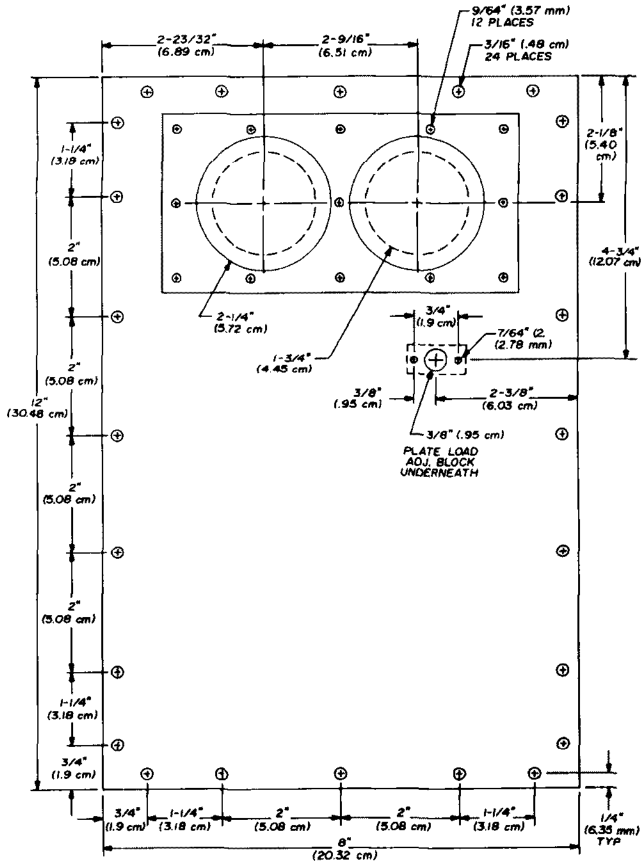




**fig. 3. Bottom view of the top plate of the two-meter kilowatt. Top plate is made from 3/32-inch (2.5mm) aluminum sheet; vent plate is made from 3/16-inch (4.5mm) thick aluminum. The dashed 1-3/4-inch (4.5cm) circle is for 4CX250R and similar tubes; the 2-1/4-inch (5.7cm) circle is for the recommended 8390 tubes. A piece of aluminum screen is cut to the same dimensions as the vent plate and mounted between the top plate and the vent plate.**

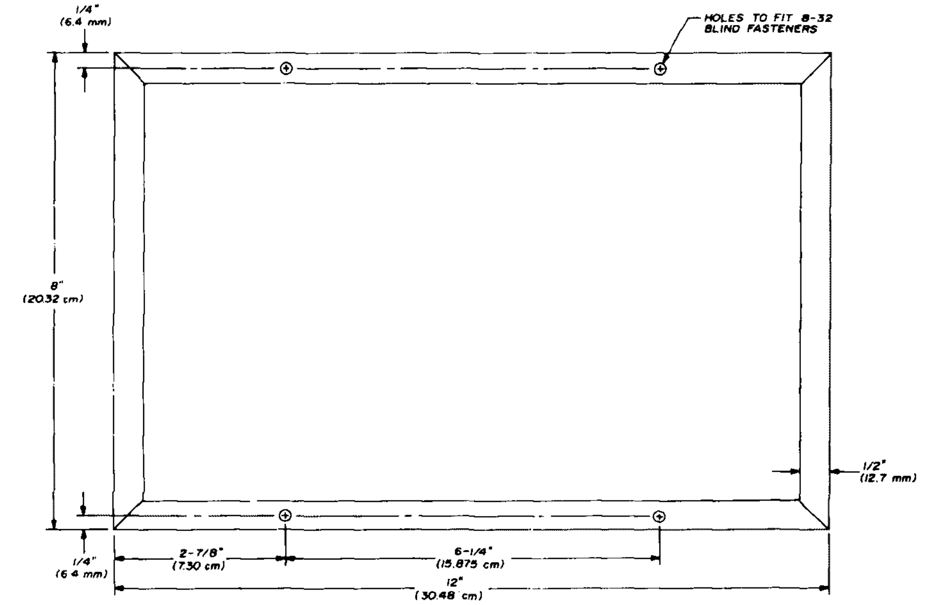



**fig. 5. Top view of the upper chassis showing the top plate mounting holes. All holes are drilled to fit 8-32 (M4) blind fasteners.**

for the chassis-mounted blower shows drilling and tapping for a Dayton 4C01 blower.

Although chassis-mounted blowers have been used successfully, vibration can be a problem. The intake plate for hose-connected blowers is drilled for a Nutone plastic fitting type 366. This fitting will accept a hose with a 2-1/4-inch (5.7cm) inside diameter. For hose-connected blowers, use a blower having 100 cfm free air rating into a 2-1/4-inch (5.7cm) aperture as a minimum. If 8930 tubes are used, the direct-mounted blower may be rated as low as 60 cfm into the same aperture.

Here are a couple of tricks to assure accurate drilling: lay the work out on masking tape which has been placed on the areas to be cut or drilled, and always use a 1/16-inch (1.5mm) pilot (starter) drill to

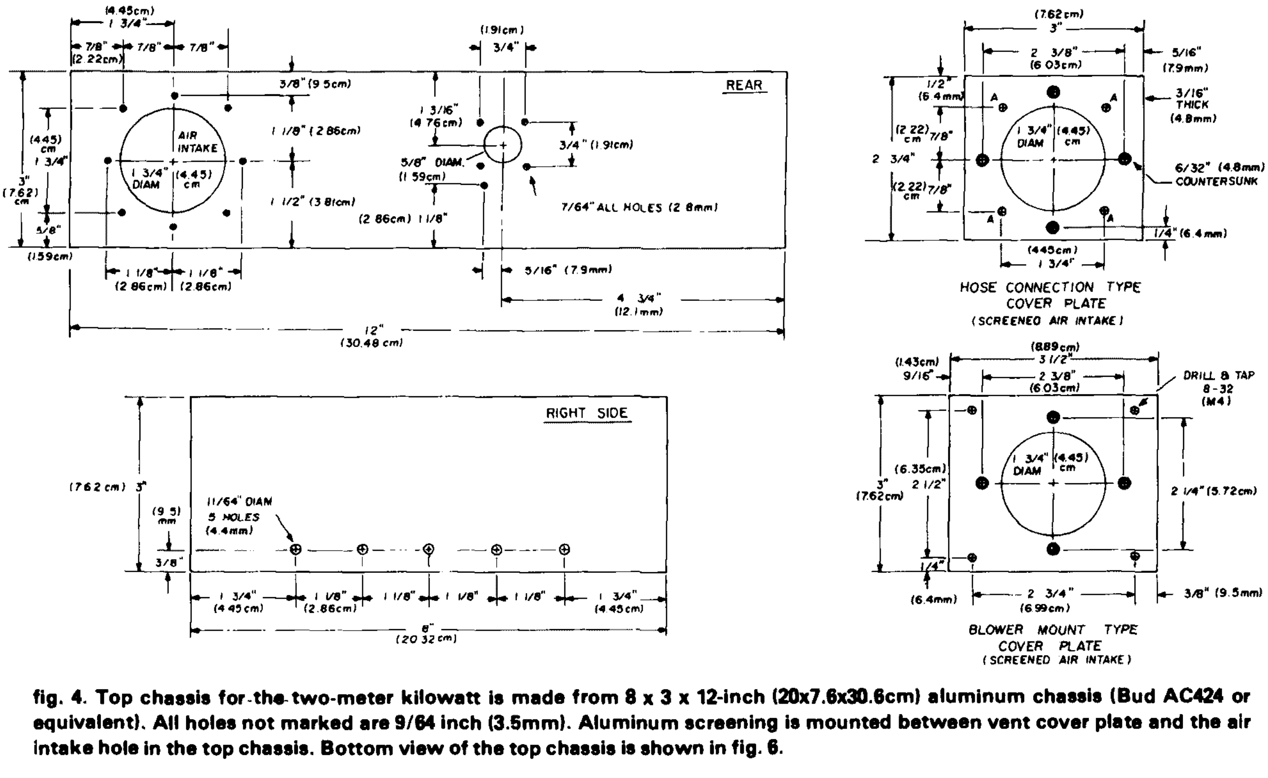



**fig. 4. Top chassis for the two-meter kilowatt is made from 8 x 3 x 12-inch (20x7.6x30.6cm) aluminum chassis (Bud AC424 or equivalent). All holes not marked are 9/64 inch (3.5mm). Aluminum screening is mounted between vent cover plate and the air intake hole in the top chassis. Bottom view of the top chassis is shown in fig. 6.**

center your holes. Although a drill press is convenient for all drilling operations, access to one is required only for the large holes through the 3/16-inch (4.5mm) thick material which requires the use of a hole saw.

To minimize alignment errors, one part with the 1/16-inch (1.5mm) pilot holes can be used as a drilling template for the matching parts. For example, use the top plate as a template for drilling the top of the upper chassis; use the vent plate as a template for the blower opening in the rear of the upper chassis.

Accuracy in drilling is essential, especially the socket holes, top plate vents, and plate line. These holes must line up nearly perfectly to assure alignment of the tubes, chimneys, and top vents.

It is best to use blind fasteners to secure the top plate to the upper chassis. Either 8-32 or 6-32 (M3.5 or M4) size is okay. The bottom plate may be fastened with self-tapping screws or blind fasteners with blind fasteners being the best choice. All other fastening is done with 6-32 or 4-40 (M3.5 or M3) screws, nuts, and lockwashers of suitable length.

The grid box is drilled and punched as shown in

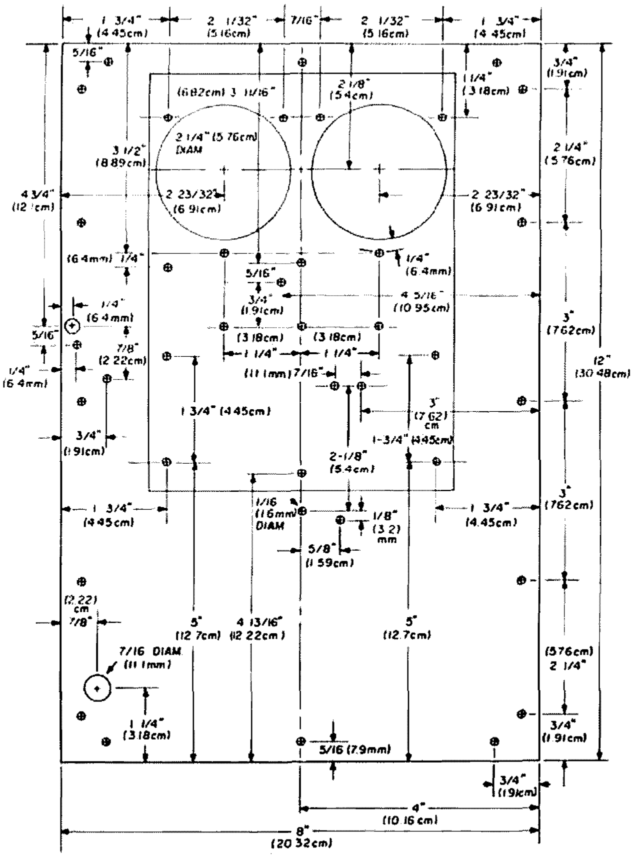
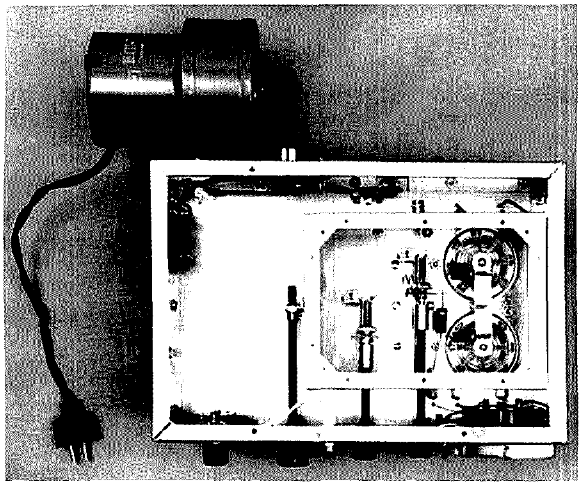



fig. 6. Bottom visw of the upper chassis showing the layout of the tube cutouts and other mounting holes. All holes not marked are 9/64-inch (3.5mm) diameter.



**Bottom view of the stripline two-meter kilowatt showing the wiring of the grid compartment.**

figs. 7 and 8. To ensure alignment of the socket holes with those in the upper chassis, the grid box pilot holes, including the pilot holes for the sockets, are drilled using the upper chassis as a template.

The top of the lower chassis is cut out as shown in fig. 9. This can be done with a nibbling tool, a small hand saw, or if you are very careful, on a table saw. The meter hole in fig. 11 is for a Calectro D10912 — a 0-1 mA meter with 100 ohms resistance. Any 0-1 mA meter not more than 2-3/4-inches (7cm) high can be used; the clearance for the meter behind the front of the chassis is 1-1/4 inch (3cm).

The holes in the rear of the lower chassis may be changed to suit your choice of power connectors. The MHV high-voltage connector (Amphenol) is recommended for the B + lead.

The vent holes in both the grid box and the lower chassis can be covered with screening by using a 1-1/8 inch (2.9cm) punch to cut the center out of the pieces punched from the 2-1/4 inch (5.7cm) socket holes. This provides a ring-shaped clamp which is drilled to match the holes in the chassis.

The bottom plate shown in fig. 8 does double duty as it is fastened to both the lower chassis and the grid box. First drill the pilot holes in the bottom plate and then, using it as a template, drill the pilot holes in the bottom of the lower chassis and the grid box. This completes the chassis work operations.

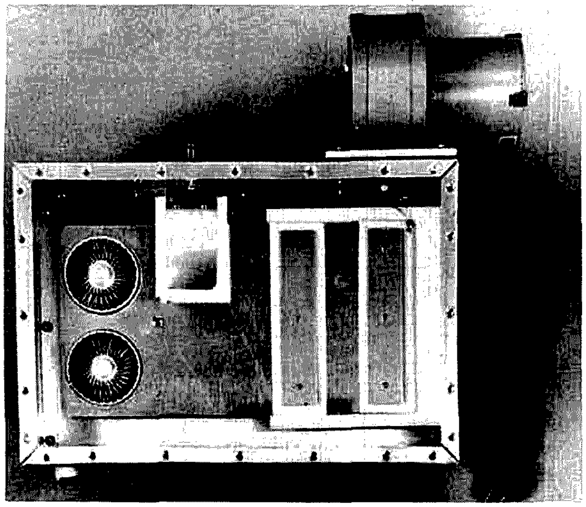
## plate line

Before starting assembly and wiring, do the cutting and drilling for the plate line, the grid line, and the output flapper. The plate line consists of two pieces of copper clamped together in a Teflon sandwich by clamping bars (fig. 12). The cutting and drilling dimensions are shown in fig. 13. Note that only two of the clamping bars are drilled and tapped.

The third bar is drilled only 9/64 inch (3.5mm).

The finger stock is soldered on the under side of the plate line; this requires a soldering tool in the 200-watt range. The finger stock is held in place, flush with the upper side of the line, with a Pyrex beaker or some other cylinder of heat-resistant material about 2 inches (5cm) in diameter (1-5/8 inch or 4cm for 4CX250s) which is wrapped with Teflon tape to provide a squeeze fit which will hold the finger stock in place while soldering. Note that the smaller piece of the copper plate line is equipped with self clinching nuts; as an alternative, brass nuts may be soldered to the copper plate. The Teflon support for the plate line at the tube end is made from 1/2-inch (13mm) diameter Teflon rod, drilled and tapped 1/2-inch (13mm) deep on each end.

**Fig. 14** shows the various components of the plate tuning and output flapper capacitors. The plate capacitor consists of two sections: the flapper and a piece of copper on the same mounting block which provides the additional capacitance needed to resonate the line. Semi-hard beryllium copper seems to make the best flapper material; it also has the advantage of taking silver plating which, while not essential, is desirable for all of the rf parts in both the plate and grid compartments of this amplifier.



**Top view of ths two-meter kilowatt showing the plate stripline, the output coupling flapper (upper center), and Teflon sandwich at the cold end of the line (right).**

Note the details for the aluminum support block in

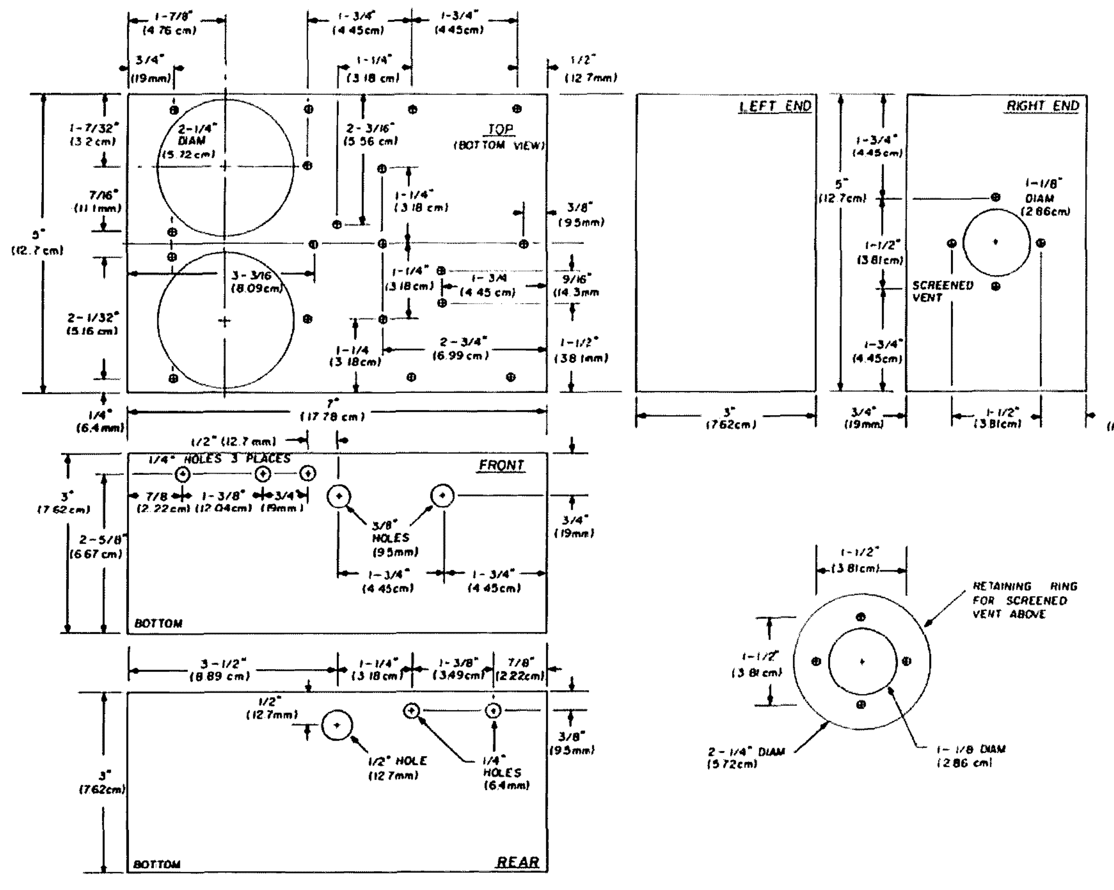
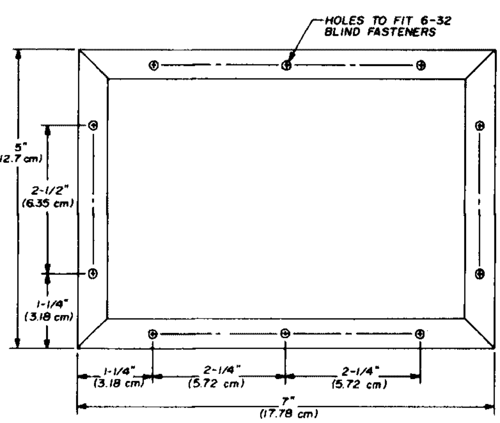



**fig. 7. Construction of the grid box for the two-meter kilowatt. The aluminum chassis measures 5 x 7 x 3 inches (12.7x17.8x7.6cm) (Bud AC429 or equivalent). A piece of aluminum screen 2-1/4 inches (5.7cm) in diameter is clamped inside the box with the retaining ring. All holes not marked are 9/64 inch (3.5mm) in diameter.**



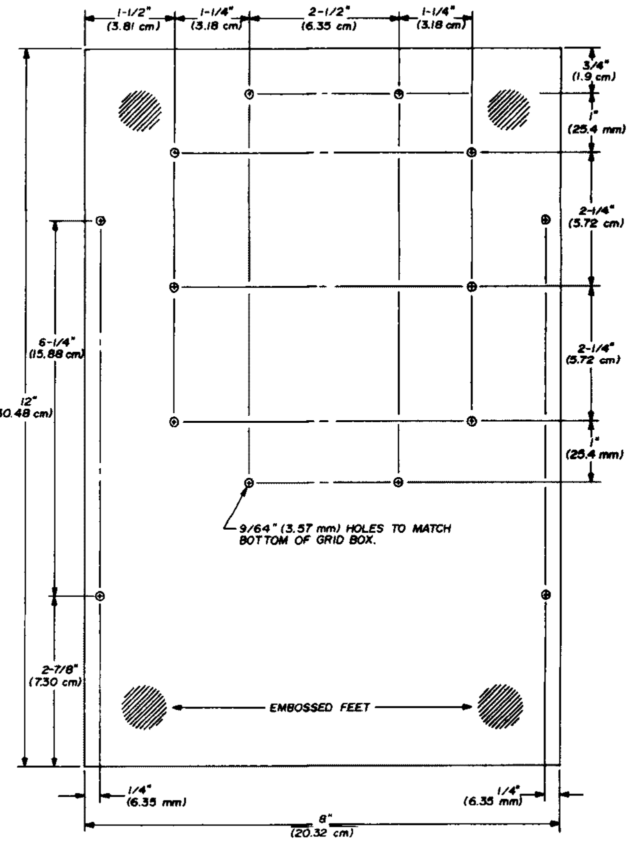




**fig. 8. Bottom of the grid box and layout of the cover plate (Bud BPA 1519 or equivalent). The holes in the bottom of the grid box are drilled using the bottom plate as a template.**

**fig. 14**; the flapper and fixed capacitor piece are mounted together with 8-32 (M4) hardware. The support itself is mounted to the chassis with 6-32 (M3.5) hardware. Dimensions for the bakelite shaft and bearing bracket for the plate flapper are also shown in **fig. 14**.

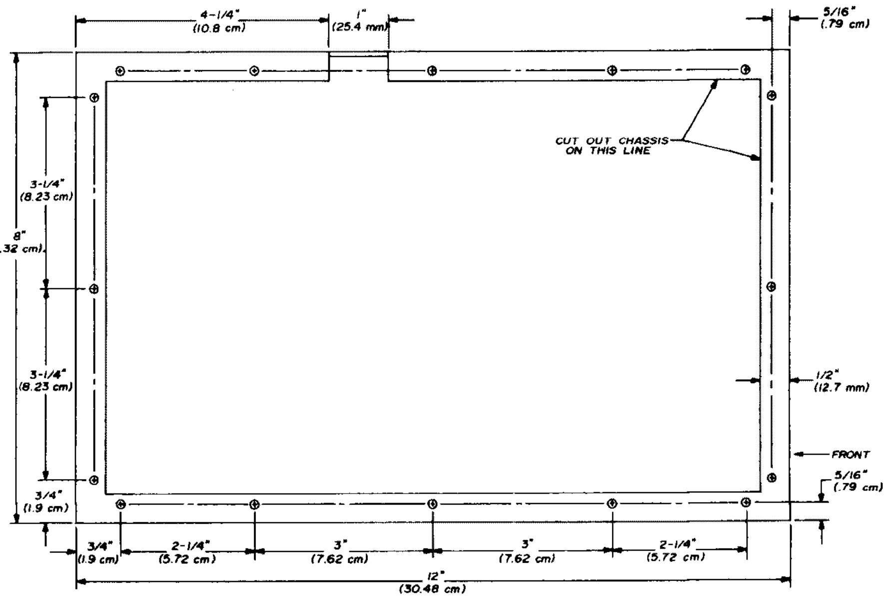



**fig. 9. Top view of the lower chassis showing the large cutout made with a nibbling tool. Chassis is a Bud AC424 or equivalent. All holes are 9/64 inch (3.5mm).**

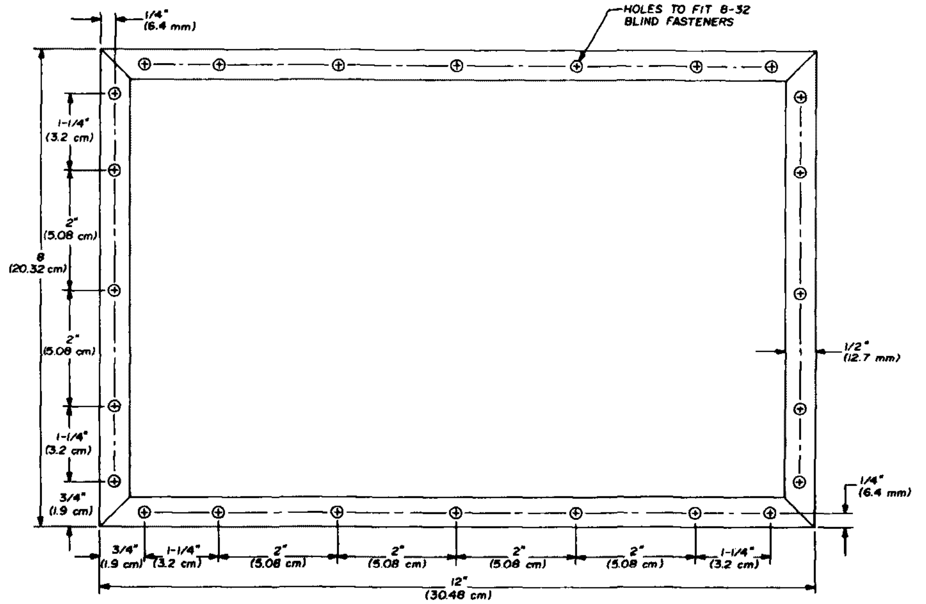


fig. 10. Bottom view of the lower chassis for the two-meter kilowatt.

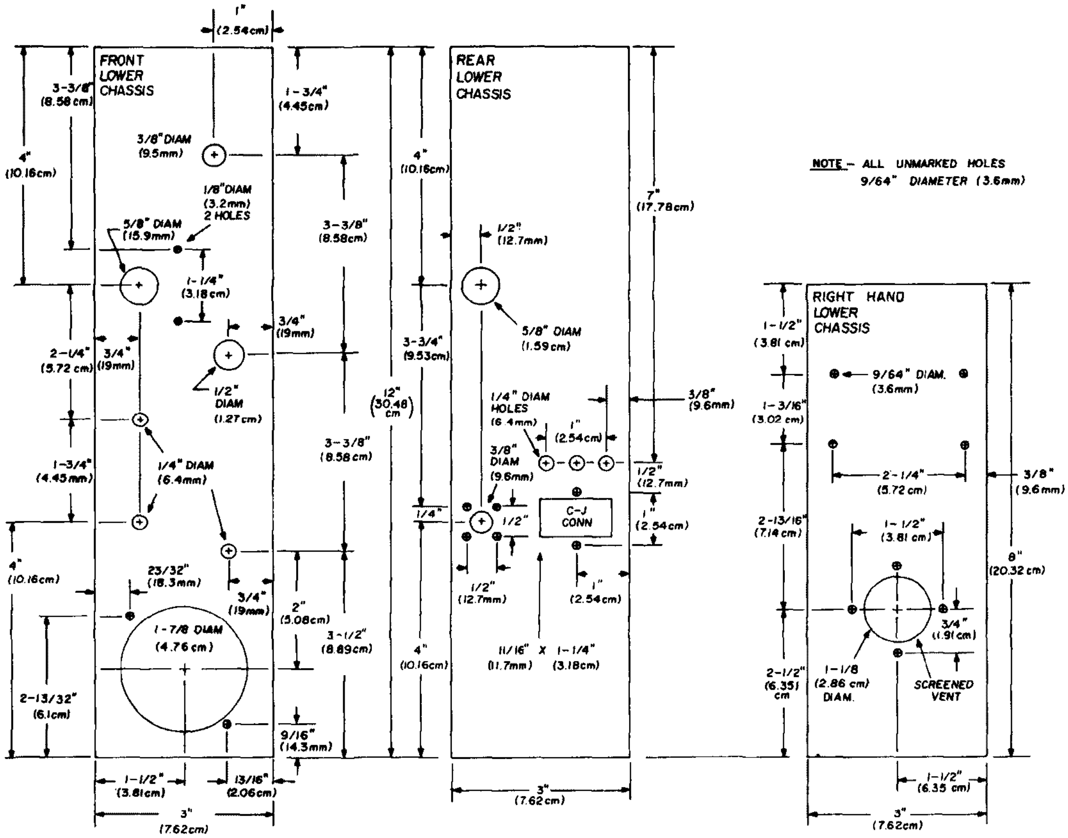


fig. 11. Layout of the front, rear, and right-hand side of the lower chassis (there are no holes in the left-hand side of the chassis). See fig. 7 for construction of the retaining ring for the screened vent. The meter hole in the front panel of the chassis is for a Calectro D10912 1 mA meter.

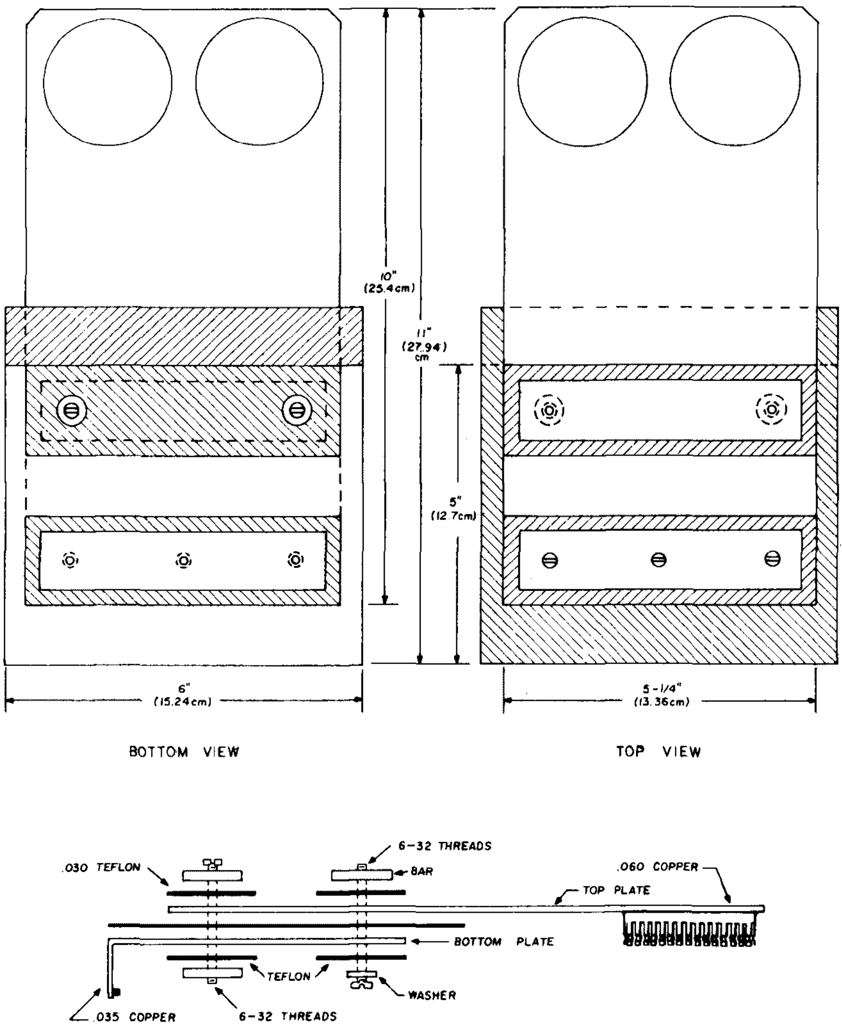



**fig. 12. Construction of the plate line for the two-meter kilowatt. Material is .060 inch (1.5mm) copper (should be silver plated for best results). The shaded areas are 0.030-inch (0.8mm) Teflon sheet. Construction of the individual parts for the plate are shown in fig. 13.**

A slot is sawed in the inner conductor of a type-N coaxial chassis connector to accept the output flapper (see **fig. 2**) which is soldered into the position shown. The two holes in the output flapper serve to mount a piece of Teflon underneath the flapper which prevents contact with the high voltage on the plate line. The dimensions of these items are shown in **fig. 14**.

Also prepare the short piece of 1/4-inch (6.5mm) diameter copper tubing for the rf sampling assembly and the piece of 1/4-inch (6.5mm) diameter Teflon rod for the rf choke form (**fig. 14**). A copper strap for connecting the grid terminals together, and mounting details for the grid butterfly capacitors and their respective shafts are shown in **fig. 15**.*

## assembly

Begin by mounting the parts and wiring the lower chassis. Connect all leads except the five leads from the grid box and the cable to the rf sampling assembly. Keep the wires formed into a bundle in the corner of the bottom of the chassis. Run in this manner, the wires will show only at the points of termination and can be laced into a cable with wire ties after they are all in place. Use a color scheme such as black for ground, green for filament, yellow for grid, and blue for screen.

*Many parts and assemblies for this two-meter amplifier are available from ARCOS, Post Office Box 546, East Greenbush, New York 12061; telephone (518) 477-4990. A price list will be furnished upon receipt of a self-addressed, stamped envelope.

Turning now to the upper chassis, install the rf sampling assembly as shown in **fig. 17**. Also install the high-voltage feedthrough capacitor at this time. Fasten the upper and lower chassis together with 6-32 (M3.5) 1/4-inch (6.5mm) long screws.

Install the bulkhead BNC input connector and the five feedthrough capacitors on the grid box. Install the grid box and the sockets, making certain that socket terminals 1 and 3 are opposite their respective feedthrough capacitors. Run the five wires from the grid box to the resistor assembly board and the RG-174/U coaxial cable to the rf sampling assembly. This completes the wiring of the lower chassis.

Mount the butterfly capacitors (see **fig. 15**). Note that there are two holes in the area where the grid tuning capacitor mounts. The capacitor assembly is

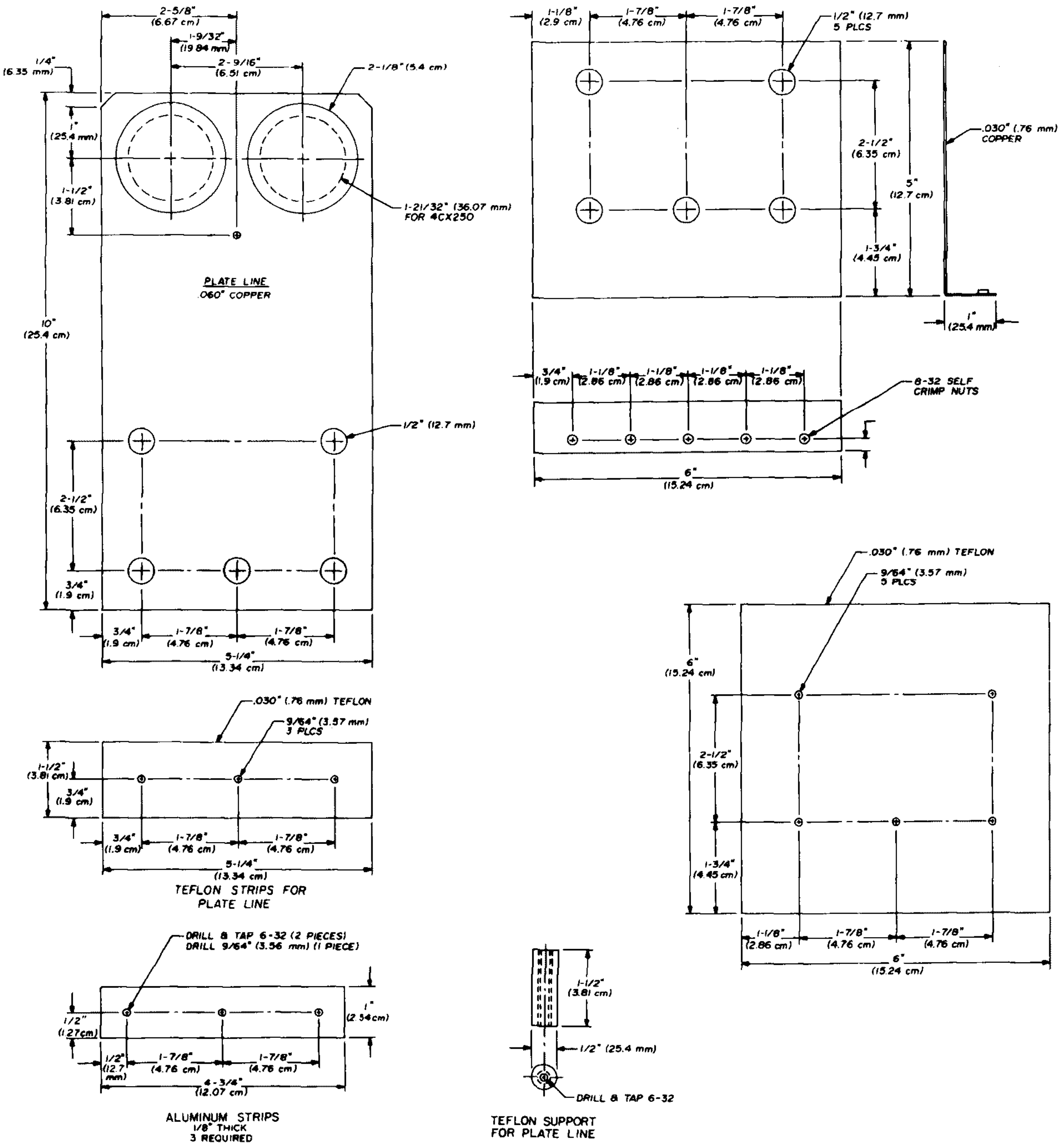



**fig. 13. Plate line parts for the two-meter stripline kilowatt.**

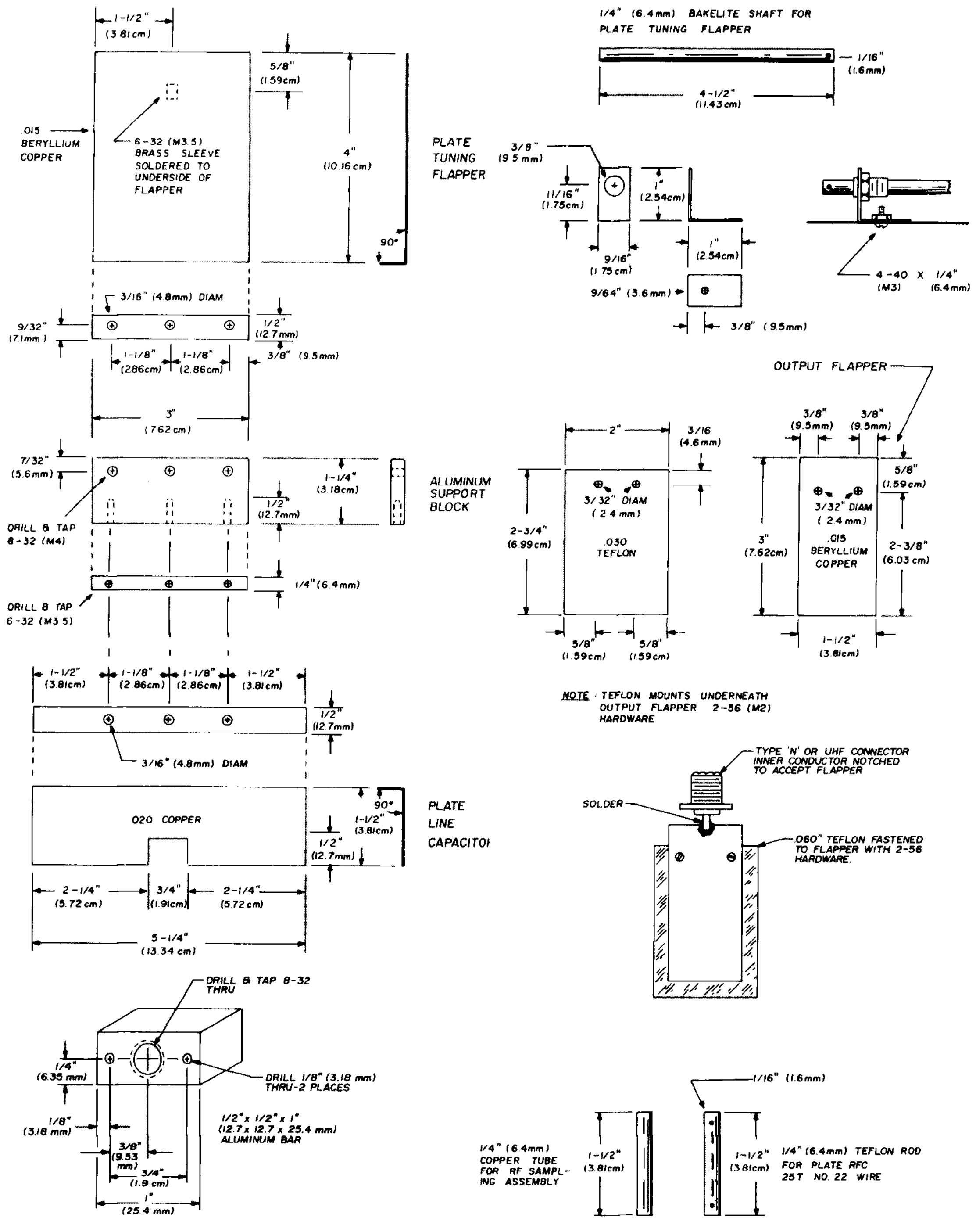



**fig. 14. Construction of the plate line tuning and output flapper capacitors.**

installed in the hole nearest the front of the chassis. Before mounting the tuning capacitor, put a spade bolt (with top cut off) in the rear hole. Tin the spade bolt. Now mount the capacitor assembly, adjusting the upper lug of the capacitor so it has some tension against the concave top of the spade bolt. Line up the coupling shaft and tighten the nut on the spade bolt holding the capacitor. Make sure the capacitor turns smoothly.

Using a 200-watt iron, carefully solder the upper lug of the butterfly capacitor to the spade lug. This is a very important connection and becomes inacces-

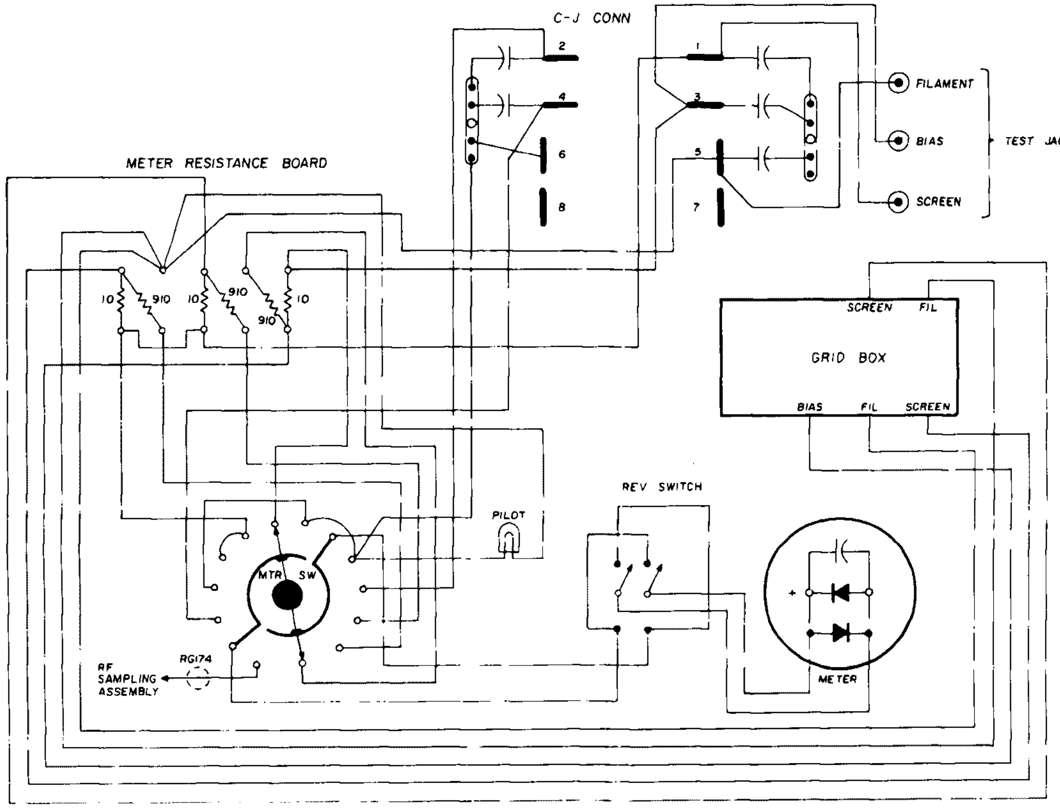



fig. 16. Wiring diagram for the two-meter kilowatt. Details of the rf sampling assembly are shown in fig. 17.

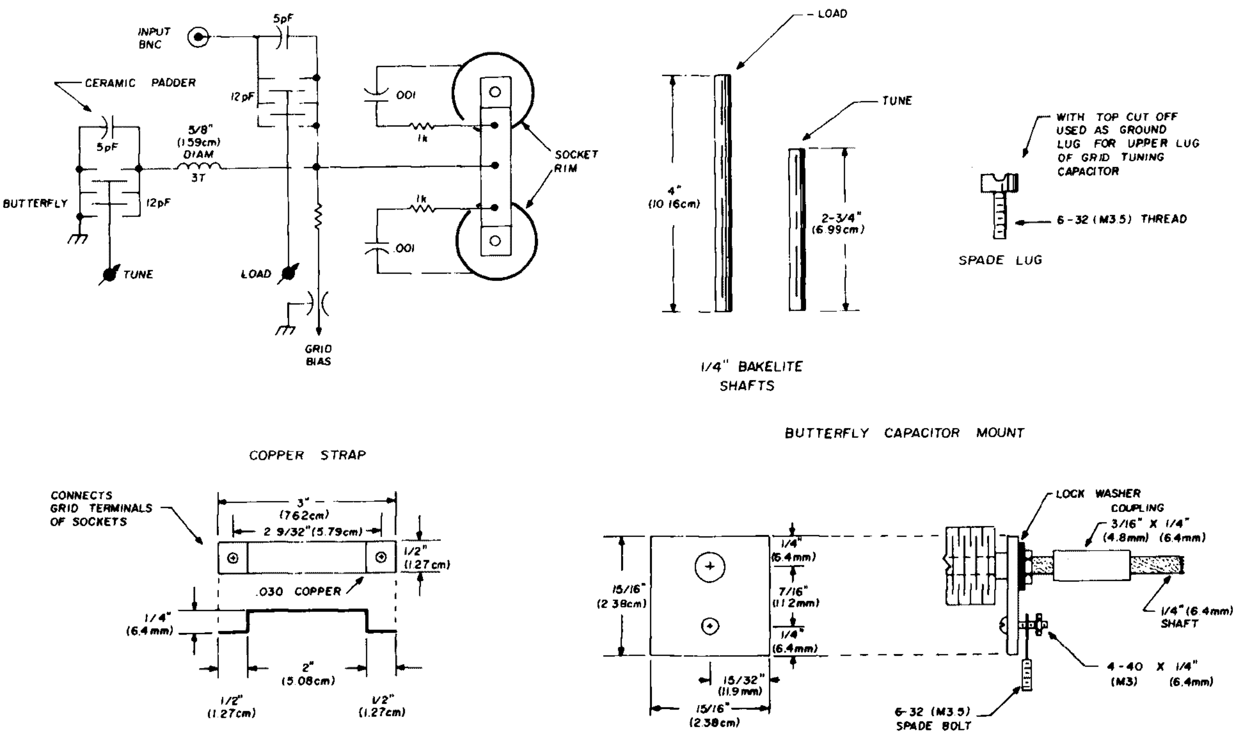



fig. 15. Grid circuit details for the two-meter stripline kilowatt.

sible later, so make sure it's a good solder joint. The other butterfly capacitor (loading) is not grounded so it is just a matter of making sure it operates freely. The balance of the grid circuit may now be installed.

Now mount the shaft bearing bracket and the bearing for the plate flapper tuning control. Install the plate tuning dial and the shaft. Make sure the shaft turns freely in the dial hub; this will facilitate placing the flapper tuning in the proper range. Install the Teflon support at the plate end of the plate line.

Assemble the two parts of the plate tuning capacitor to the aluminum support block and bolt the support block to the chassis. (The fishline control should be made fast to the plate tuning flapper before mounting the flapper assembly to the chassis, and the plate flapper positioned to about 1-3/4 inch (4.5cm) above the chassis.) Assemble the plate line (see fig. 12) and mount it to the chassis. Before tightening up the Teflon sandwich make sure that the large piece of Teflon between the plates is centered in the mounting holes.

Install the rf choke and the output flapper (which should be bent up to within 1/4 inch or 6.5mm of the top of the chassis). Position the rf sampling capacitor about 1/8 inch (3mm) away from the output flapper

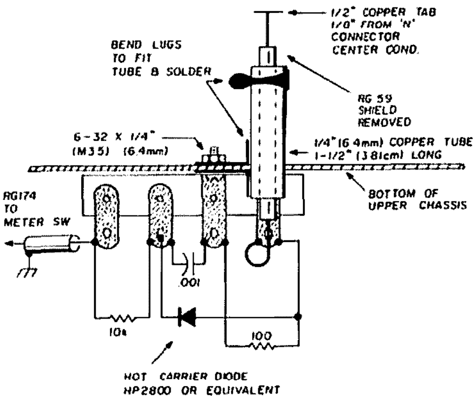
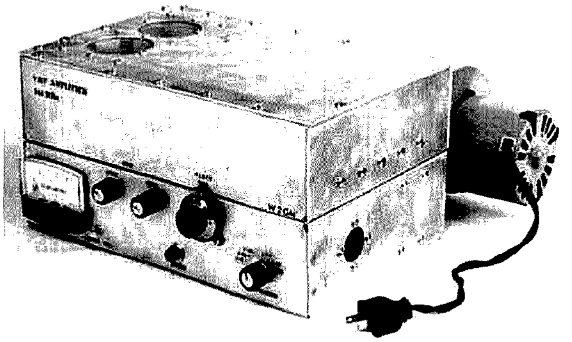



**fig. 17. Side view of the rf sampling assembly. Adjust the position of the copper tab for approximately 0.6 mA on the meter at 600 watts output.**

and type-N connector. Now assemble the top plate vent and the loading adjustment block. Fasten the top plate and bottom cover to complete the assembly of the amplifier.

## power supply

A power supply suitable for this amplifier is shown in fig. 21. The voltage-doubling circuit offers a 1000 volt source for the screen dropping resistor. Since no grid current will flow with linear operation, the bias supply can be the simple zener regulated type



**The entire amplifier is assembled in two mated chassis.**

shown. The protective features of this supply include a high voltage fuse and a diode protective resistor in the transformer secondary lead. A delay tube maintains cutoff bias until the tubes warm up. The filament voltage is 8 volts at the transformer winding and offers ample control of the voltage at the sockets with the adjustable 1/2-ohm series resistor.

This power supply will deliver 2000 volts at 500 mA with a no-load voltage of 2200 volts. At 1 ampere the output voltage drops to 1850 volts. With a transformer weight of 30 pounds (13kg) the total weight of the unit is only 45 pounds (20kg). This is quite a relief from the 80 to 100 pound (35-45kg) power supplies of the past.

## test and check out

An inexpensive dummy load for a high power vhf amplifier can be set up with 100 feet or 30 meters (or more) of RG-8/U with a Heath Cantenna at the end. This will stand up on 144 MHz for about ten minutes or so at 600 watts output before the cooling oil in the dummy load starts to boil. The driver for this amplifier *must* have a adjustable output control unless it is capable of less than 5 watts maximum output because the amplifier is very power sensitive; it is not possible to adjust it properly if it is overdriven.

After connecting the amplifier to the power supply

**table 1. Typical operation of the two-meter stripline kilowatt using 8930 ceramic power tetrodes (grid bias = −77 volts, screen supply = 409 volts, idling current = 100 mA).**

| drive power | grid current | screen 1 current | screen 2 current | power input | power output |
|---|---|---|---|---|---|
| 2 W | 0 | −6 mA | −6 mA | 800 W | 200 W |
| 4 W | 0 | −8 mA | −10 mA | 1000 W | 400 W |
| 8 W | −1 mA | +1 mA | +2 mA | 1360 W | 830 W |

Plate voltage during these tests ranged from 2100 volts at 200 mA idling current, to 1700 volts at 830 watts output.

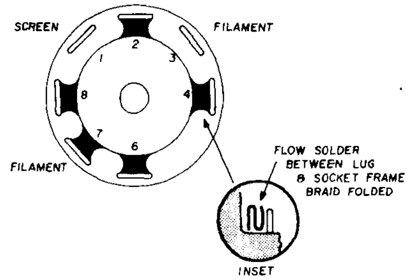



**fig. 18. Grounding for the Eimac 620A socket. The 630A socket requires this modification only on terminal 7.**

and making the usual checks of filament voltage, bias, screen voltage, and blower operation, establish an idling current of about 150 mA for initial tests. Applying a watt or so of excitation, adjust the grid circuit controls for a rise in plate current. Then resonate the plate circuit by observing power output. If the plate circuit will not resonate, change the range of the plate tuning flapper controls until the dial is midrange for your chosen operating frequency.

The next step is to increase drive until the amplifier is at about 400 watts or so output with the loading screw about 1/8 inch (3mm) above the top plate. Now the grid circuit controls should be set for minimum swr toward the driving source. The reverse power will drop to an unreadable value. Once grid tuning is established, adjust the load control on the top of the amplifier for a compromise between maximum output and minimum plate current at an output level of 600 watts. Keep the loading on the heavy side.

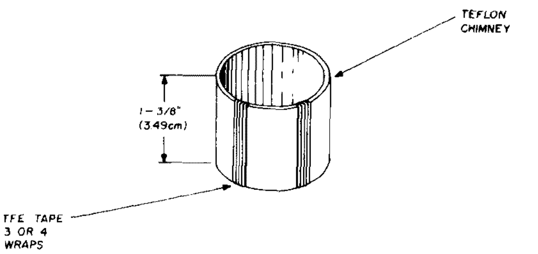



**fig. 20. Tube chimneys for the two-meter kilowatt are made from 0.010-inch (0.25mm) Teflon, 1-3/8-inches (35mm) wide. Roll the Teflon for a tight fit in the top vent plate; secure with 1/4-inch (6.5mm) Teflon tape wrapped vertically around the roll. 12 inches (30cm) of Teflon is required for 4CX250s; 24 inches (60cm) is required for 8390s.**

Once the proper setting for the load control is found, there should be no reason to change it unless you wish to operate the amplifier at a lower power level or with different voltatges. Having set up the initial operating conditions, further tests and adjustments can be made for best linearity.

Check for blower operation each time the amplifier is turned on. If the air supply should fail, the solder on the plate line will melt and the finger stock usually springs out, grounding the plate supply and operating the breaker within about 30 seconds. The tubes survive but it is a messy job to repair the plate line.

## safety

Like any other piece of radio equipment operating at high voltages, this innocent looking aluminum box can be a killer — it is absolutely unforgiving of careless moves. Just to repeat the safety rules, disconnect the B+ line before taking covers off. Don't operate the amplifier with the covers off. If you must take a reading inside with the cover off, disconnect the power, connect the meter with well-insulated leads, stand back and take the reading *after* you have put the power plug back in. One hand in the pocket while you are testing is the time-tested rule

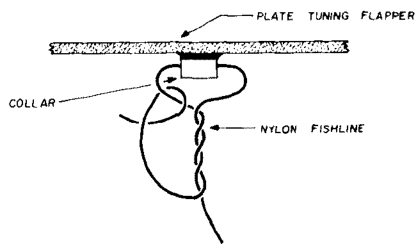


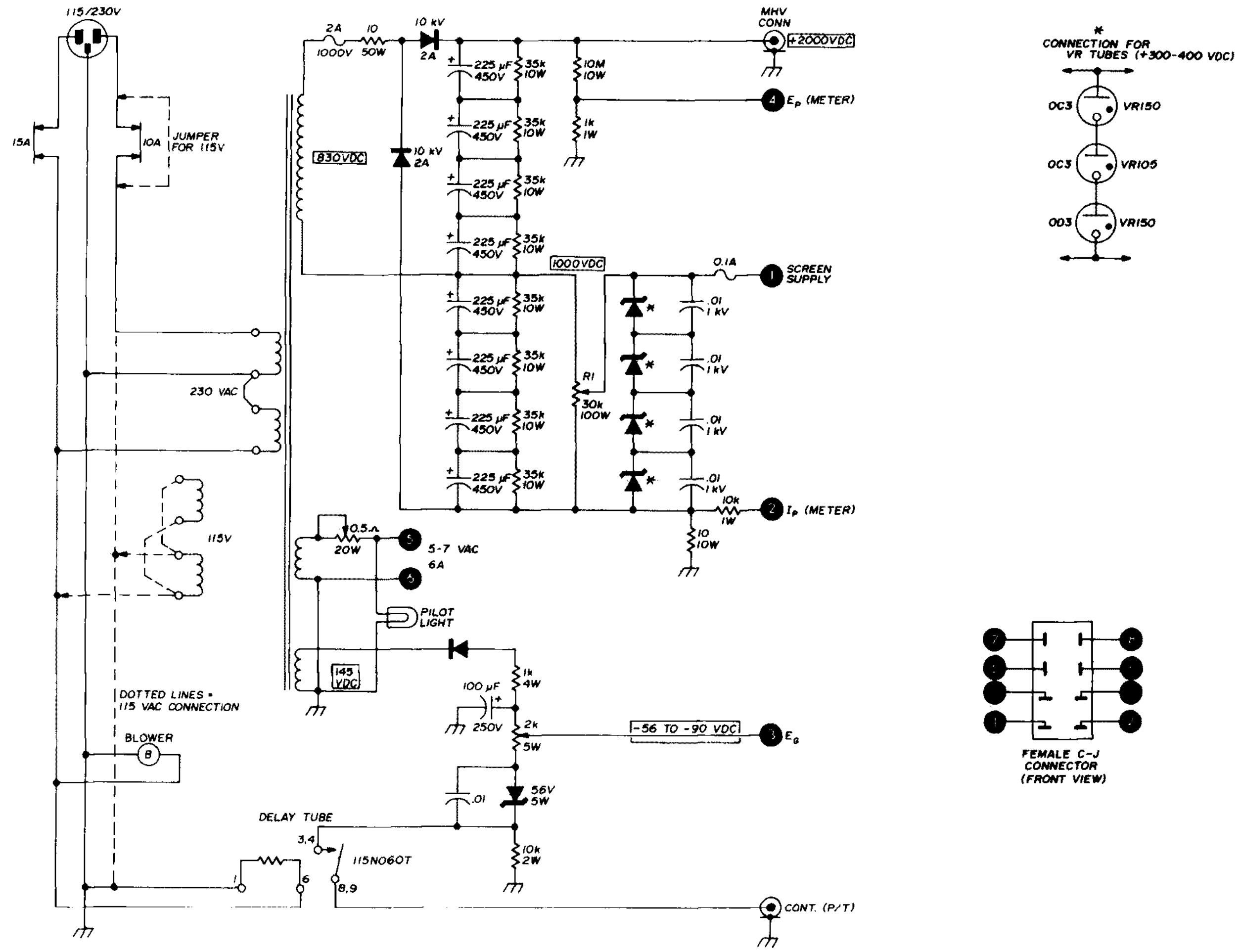

**fig. 19. Knot for fastening nylon fishline to the collar on the plate tuning flapper, and to the tuning shaft. Pull the knot tight to prevent unraveling.**

for staying alive. Another precaution in dealing with this level of rf power is to stay out of the path of the rf radiated from the antenna.

## operation

To adjust the amplifier for linear operation, perhaps the simplest setup is to use a directional wattmeter at both the input and the output. Set the idling current at 200 mA. Using the test values of 2, 4, and 8 watts drive, (see **table 1**)adjust the loading control and the plate tuning control for best linearity at the lowest value of peak plate current. When the proper adjustment is reached the power output will be approximately 200, 400, and 800 watts, respectively. These tests can be made at the bench and the adjustments will be valid for a 50-ohm antenna system with low swr.

For CW operation the bias can be set higher for an idling current of 100 mA or less, and the loading control and plate tuning optimized for best output at the least plate current. At 1 kilowatt input, 600 watts output is the objective. About 5 watts drive is required.




**Voltage:** no load, 2300 V; 500 mA, 2000 V; 1 A, 1900 V
**Bias:** −55 to −95 volts (no grid current)

**Standby bias:** −150 volts (P/T contacts open)
**Weight:** 45 pounds (20 kg)

**fig. 21. Recommended power supply for the two-meter stripline kilowatt is rated at 1 kilowatt continuous duty, 2 kW PEP. Adjust R1 (30k, 100W) for 500 volts at the screen terminal with the zener diodes (or VR tubes) disconnected. Readings in boxes were taken at 1 kilowatt output.**

The temperature of the exhaust ports can be checked by setting a candy thermometer on top of the amplifier in a nearly vertical position (put a pair of stiff wire legs on it). This does not give a reading of the temperatures at the tube seals, but is a relative indicator. Based on many hours of testing these amplifiers, a suggested limit for the exhaust air temperature read in this manner is 200°F (95°C). The tubes will not fail under continuous operation at this temperature limit. The 8930s will stay well within this level but the 4CX250s will tend to exceed it.

The filament voltage for the two-meter amplifier has been set at 6 volts on the assumption that the effects requiring 5.5 volts on the filaments at 432 MHz are not as significant at 144 MHz.

The grid current should be at zero or barely negative for AB1 operation. Screen current will go negative (read with the meter reverse switch) to 10 mA more or less. It may become barely positive on modulation peaks. The screen voltage is regulated at 350 to 400 volts.

Once you have this amplifier on the air, properly adjusted, you will find that no further attention is required except perhaps an occasional tweaking of the plate tuning to match the antenna.

## references


1. Richard T. Knadle, Jr., K2RIW, "A Stripline Kilowatt for 432 MHz," *QST*, April, 1972, page 49; May, 1972, page 59.
2. Steve Gross, W9OJI, "A Parallel 4CX250B Amplifier for 144 MHz," *QST*, May, 1975, page 11.


**ham radio**

# how to improve the accuracy of your frequency counter

Circuit details for a phased-lock time base and sensitive wideband probe that will improve the accuracy and usefulness of your frequency counter

**Most electronic equipment** requires a warmup period to allow them to stabilize and become ready for use, and the solid-state frequency counter is no exception; many instruction manuals caution that no critical frequency measurements should be made until the counter has warmed up for 30 minutes or more. If you have an accurate frequency standard, you can use the output of this standard to eliminate any warmup drift and to greatly improve the accuracy of your frequency measurements.

Some years ago, being interested in accurate frequency measurement, I built an oven mounted, temperature controlled 100 kHz crystal oscillator, divided the frequency to 60 Hz, and used the amplified output to run a synchronous electric clock with a sweep second hand. I can set this clock to the exact time as broadcast by WWV and, with a little care, the crystal frequency can be adjusted so the clock keeps exact time. If the clock stays within one second of WWV time for a 12-day period, this means that the 100-kHz oscillator is accurate to one part in $10^6$; if it can be held to one second in 4 months, this represents an accuracy of one part in $10^7$, of or a little more than one cycle per second in 14 million. This accuracy can be easily translated to the time-base oscillator in your frequency counter.

The construction details of the 100-kHz standard

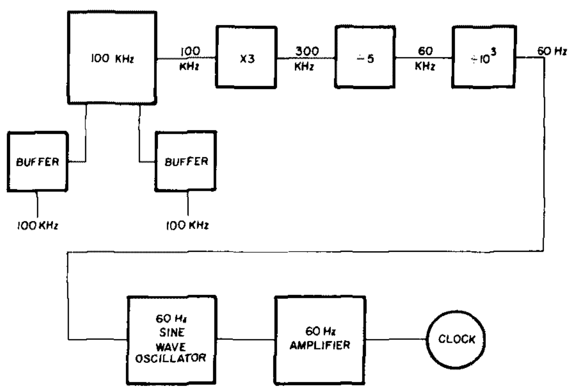



**fig. 1. Block diagram of the 100-kHz frequency standard used by W1RF to drive a clock. This same basic circuit can be used to provide excellent time-base stability when used in a frequency counter.**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are beyond the scope of this article, but the block diagram of **fig. 1** shows the basic idea. The schematic of the crystal oscillator and associated countdown circuitry in my counter is shown in **fig. 2**. I used a 4-MHz crystal in the oscillator because I had a number of old 75-meter phoneband crystals on hand, and it was a simple matter, with a little careful grinding to move one up to 4 MHz. With a different division ratio, you can use a different frequency crystal.

The total capacitance in series with the crystal, necessary for setting it precisely to 4 MHz, consists of a fixed capacitor in parallel with a variable capacitor in parallel with a voltage variable capacitor, a varicap diode, or in my case, a silicon rectifier diode. The control voltage for this diode is developed by a Motorola MC4044P phase frequency detector and associated amplifier and filter. As shown, the 4-MHz signal is divided by four in a 7473 IC, and further divided by 10 in a 7490; the output is at 100 kHz. Part of this output is taken off through a 330-ohm resistor and the main output being further divided by additional 7490s (down to 1 Hz in my counter).

The output from the 100-kHz buffer is less than 0.5 volt, so I had to provide additional gain with Q1 so U2A would saturate properly. There are hundreds of transistor types which can be used for Q1, Q2, and Q3; the ones I used came from surplus board assemblies so the type numbers are unknown.

A good way to select proper component values is to put an oscilloscope probe on pin 1 of the MC4044P and adjust C3 and R1 for a symmetrical square wave with clean leading and trailing edges. Capacitor C3 in my unit is 68 pF; R1 is 300k. These values will vary, depending on the types of NPN transistors you use. The 470-pF capacitor from the collector of Q2 to ground was required in my circuit to remove a high-frequency oscillation appearing on the leading edge of the square wave. Gates U2B, U2C, and transistor Q3 form a lock indicator circuit; when the 4-MHz crystal oscillator is locked to the 100-kHz standard,

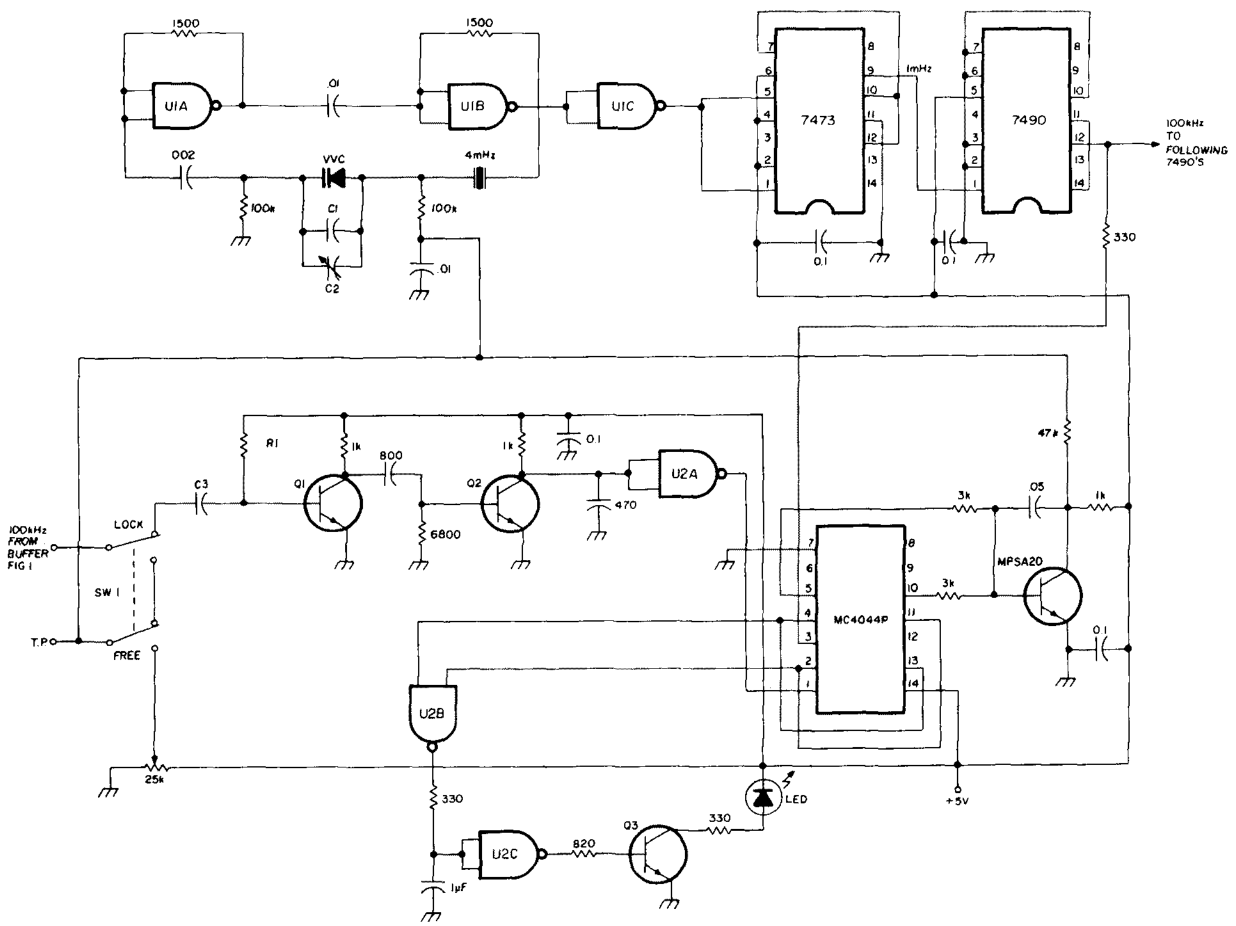


**fig. 2. Circuit diagram of the phase-locked frequency standard. The author used a 4-MHz crystal, but other frequencies are possible if the division ratio is changed accordingly. The oscillator circuit is phased locked to the output of a high-stability 100-kHz frequency standard (see fig. 1). U1 and U2 are SN7400s.**

the LED will light. Don't forget that the LED is a diode and will light only when properly connected into the circuit.

Switch S1 is helpful in initially setting up the circuit, and allows the counter to be used for normal accuracy readout anytime it is not necessary or desirable to have it phased locked to the frequency standard. With S1 in the *free* position and a vtvm connected to test point TP, set the 25k pot so the meter reads about 3 volts. With C2 set at about the mid-capacitance position, select a value for C1 such

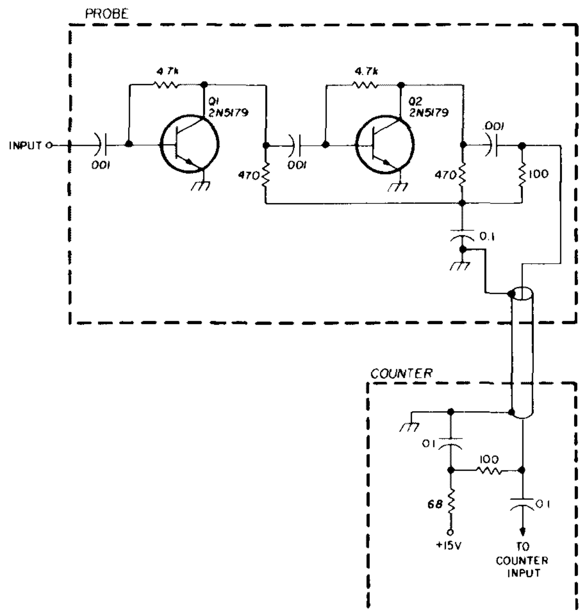


**fig. 3. Rf proba for frequency counters which greatly increases sensitivity at the higher frequencies. Power for the probe is provided through the coaxial input cable.**

that the crystal oscillator is very nearly on the correct frequency. In my circuit C2 is a 5-35 pF trimmer set for about 20 pF, and C1 is 20 pF.

With the 100-kHz standard connected and S1 in the *lock* position, the LED should light and the vtvm should read somewhere between 1 and 5 volts. C2 should now be readjusted for a vtvm reading of about 3 volts. If the counter is now connected to read out the frequency of a stable external oscillator, note that varying C2 will not change the readout but will only vary the control voltage. If the LED does not light and no lock is obtained regardless of the setting of C2, the vtvm will probably read 5 volts; this indicates that the crystal is too far off frequency to be brought in by the varicaps, or the varicaps do not have enough range.

In my counter, running free and measuring a 5-MHz external oscillator, varying the 25k pot from zero to 5 volts changes the readout from 4 999 980 to 5 000 184 Hz. Using a silicon power diode as a varicap is somewhat questionable as some diodes have much greater capacitance range than others of the same type. If you run into trouble, a Motorola MV-series varicap should be tried (such as MV2110).

With the time-base oscillator locked to a good frequency standard you can be sure that both the accuracy and the stability of your counter is greatly improved. Some years ago, after completing a similar phase-locked counter, I set up the counter to read out the frequency of my transmitter exciter. One evening I tuned in on the ARRL Frequency Measuring Test. Due to skip conditions I could hear only two of the transmitted frequencies, but it was a simple matter to zero beat W1AW with my exciter and read out the frequencies. When the actual transmitted frequencies were published a month or two later, I was pleased to note that of the two readings I submitted, one was precisely as published and the other was 1 Hz off! If you want to get to the top of the list in a Frequency Measuring Test, all you need is a phase-locked oscillator in your counter, and a little luck!

Another useful accessory to my counter is a probe based on the broadband amplifier described previously in *ham radio*.[1] One trouble you run into when measuring high frequency signal in low-power circuits is the difficulty of obtaining sufficient voltage to actuate the counter. If you connect the circuit under test to the counter with coaxial cable, capacitance of two or three feet (1 meter) of cable can produce a relatively low impedance, and an unwanted load. The broadband amplifier appeared to be a solution to the problem, but I disliked hooking up the amplifier and associated power supply every time I wanted to use it so I modified it slightly and built it into a probe. The simple modification shown in **fig. 3** allows the amplifier to be powered over the same coax that carries the amplified input.

If you have 15 volts at about 20 mA available in your counter to power the amplifier, the probe is very convenient to use. If you are handy with a lathe the probe can be made to look quite professional; mine is 3/4 inch (1.9cm) in diameter and 5 inches (12.7cm) long. With the probe I need only 5 millivolts at 40 MHz, 2 millivolts at 2 MHz to operate my counter. Since the center conductor of the coax connector on the counter is carrying 15 volts, be sure to use an external isolating capacitor if a direct connection is made to the circuit under test.

## reference


1. Randall Rhea, WB4KSS, "General-Purpose Wideband RF Amplifier," *ham radio*, April, 1975, page 58.


**ham radio**

# syllabic vox system

## for the Collins S-Line

### Full break-in capability is featured with the Collins S-Line, using an array of switching transistors

**Rapid, silent switching** of transmitter-receiver functions will always improve a station's "on-the-air" effectiveness. By eliminating the slow dynamic operation of mechanical relays, caused in part by conventional vox systems, the operator may then enjoy a higher quality of communications. There never will be a situation of *doubling* with another station because you can be interrupted at any time with the help of this high-speed switching network.

Recently Ray Hitchcock, W6RM,[1] updated my earlier vox system[2] by using logic circuitry which offered many advantages over the older style. This article concerns itself with the conversion of his syllabic vox for the Drake equipment to the Collins S-Line. When he and I are using ssb, the conversation is essentially the same as using the telephone. If a visiting operator were to use the station, he could use the equipment in a normal fashion and not be aware of any modifications except that he would be hearing between his transmitted syllables and code elements. There are no external devices which require adjustment nor is defacement of the equipment necessary. The 5 Vdc power supply (**fig. 1**), is housed in the speaker cabinet. The original vox controls (*vox gain* and *anti-vox*) perform the same basic functions as previously. The *vox delay* circuit is rendered inoperative by this modification.

Unless two separate antennas are used, the most critical portion of the syllabic vox system is at the antenna change-over location. W6RM has refined the diode antenna switch into a well designed, trouble-free device and suggests the appropriate name, "Diode-Biased Antenna Gate." The actual switching takes place 200 μs before the rf appears and is accomplished by forward biasing the diode with a dc voltage. No rf rectification occurs as in former T/R switches.

Directions for intial setup and the logic theory of the system are given in W6RM's paper; you are urged to consult this excellent article. One serious problem arose with the 32S-3 conversion: an intolerable audio oscillation occurred with only a moderate degree of vox gain. The vox was thrown into activation making correction of this defect mandatory.

You will note differences around the LM3900 (U1A), see **fig.** 2, as compared to the schematic diagram in W6RM's article. The input to the LM3900 comprises a highpass filter, and the 74122 (U2) delays vox activation for 2.1 milliseconds.

The syllabic vox amounts to a very high speed vox without the use of relays. In the original relay circuitry each set of contacts opens and closes a single circuit. In the new vox, each set of contacts has been replaced by a switching transistor. The circuits are coordinated so that each syllable or code element controls the transmitter-receiver functions. All switching is done in the proper sequence.

### construction

The technique that I followed was point-to-point hand wiring with number 22 AWG (0.6mm) tinned wire. Teflon sleeving is used to cover this wire where required. To facilitate the connections to the innumerable power and ground points, the periphery of the usable portion of the board is enclosed by two wire loops, each terminating at the rear of the board. A solder lug is bolted to the board at each termination point (see photograph). Solder lugs are bolted to

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each point where interconnecting cables are attached. The 6.3 Vac jack, at the left rear, supplies the power to the 5 Vdc source (the spare jack may be used for the 5 volt input).

For the most part, the general layout follows that of the schematic diagram. The 2-mH choke is fastened to the rear of the bfo using one of the already available screws.

automatically places one or the other of the gates into the circuit. The relays play no part in the rf operation.

## 32S-3 modifications

Upon inspection of the chassis when placed upside down, the vox relay is at the left center. The drawing (**fig.** 3) will help identify the contacts of the relay. Attached to the front panel, in front of this relay, is the *emission* switch (S8). Wafer D is the wafer closest to the front panel. At its top (when upside down) is terminal 12 of wafer D with a connection to terminal 12 of wafer C. Terminal 11 is unused. Terminals 10, 9, 8 and 7 continue to the left.

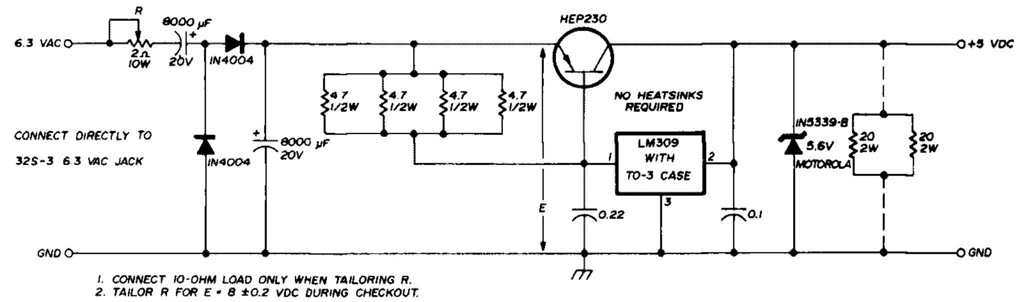



**fig. 1. Power supply for the integrated circuits. The 6.3 Vac is obtained from the connector on the rear panel of the transmitter.**

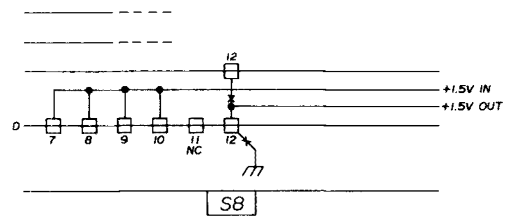
A second circuit board was mounted at the opposite end of the chassis as shown in the photograph. It holds the high-voltage transistors and is convenient to the high-voltage chassis connections. The 100k resistor becomes so hot that it must be elevated above the board for ventilation. All other components are barely warm to the touch after a prolonged operating period.

## diode-biased antenna gate

A metal box measuring 1-1/2 x 2 x 2-3/4 inches (38x51x70mm) houses the gate. Two coax fittings are mounted at the ends and two phono jacks are on top of the box. The coax fittings are connected by RG-8/U coaxial cable cut 2 inches (51mm) longer than necessary. The vinyl is removed, leaving the center 1 inch (25mm) of the braid unexposed. The wire sheathing is skinned back over the center and the wire is cut to fit into the notches after removing a bit of the foam, center insulation. Solder the wire. At one end solder one terminal of the 50 pF capacitor after covering the wire with Teflon. Then cover both bare ends with tape so that when the shielding is brought forward it may be soldered to lugs in the screws holding the fittings and thus provide proper rf shielding. The remaining parts for the gate are easily placed within the box.

The Collins coax fitting did not lend itself well to rigid mounting, so the shortest possible length of coax was used. The entire length, including fittings, was only 4 inches (100mm).

The two gates shown by W6RM are to be used when the linear amplifier is placed at a distance from the exciter. Note that the relay in each box




**1.** At section 2 of the vox relay disconnect the lead to the antenna relay from the normally open, non-movable contact.

**2.** At section 4 of the vox relay disconnect the lead to ground at the movable contact (later this lead will be used to ground the several cable shields which converge in this region). Next, connect the circuit board extension lead labelled "Mute Line" to the movable contact.

**3.** At section 3 of the vox relay connect the circuit board extension lead labelled "Mixer Hold" to the movable contact.

**4.** At section 1 of the vox relay connect the circuit board extension lead labelled "+270 V" (non-movable contact).

**5.** At the movable contact of section 1 of the relay connect the circuit board extension lead labelled "+275 V".

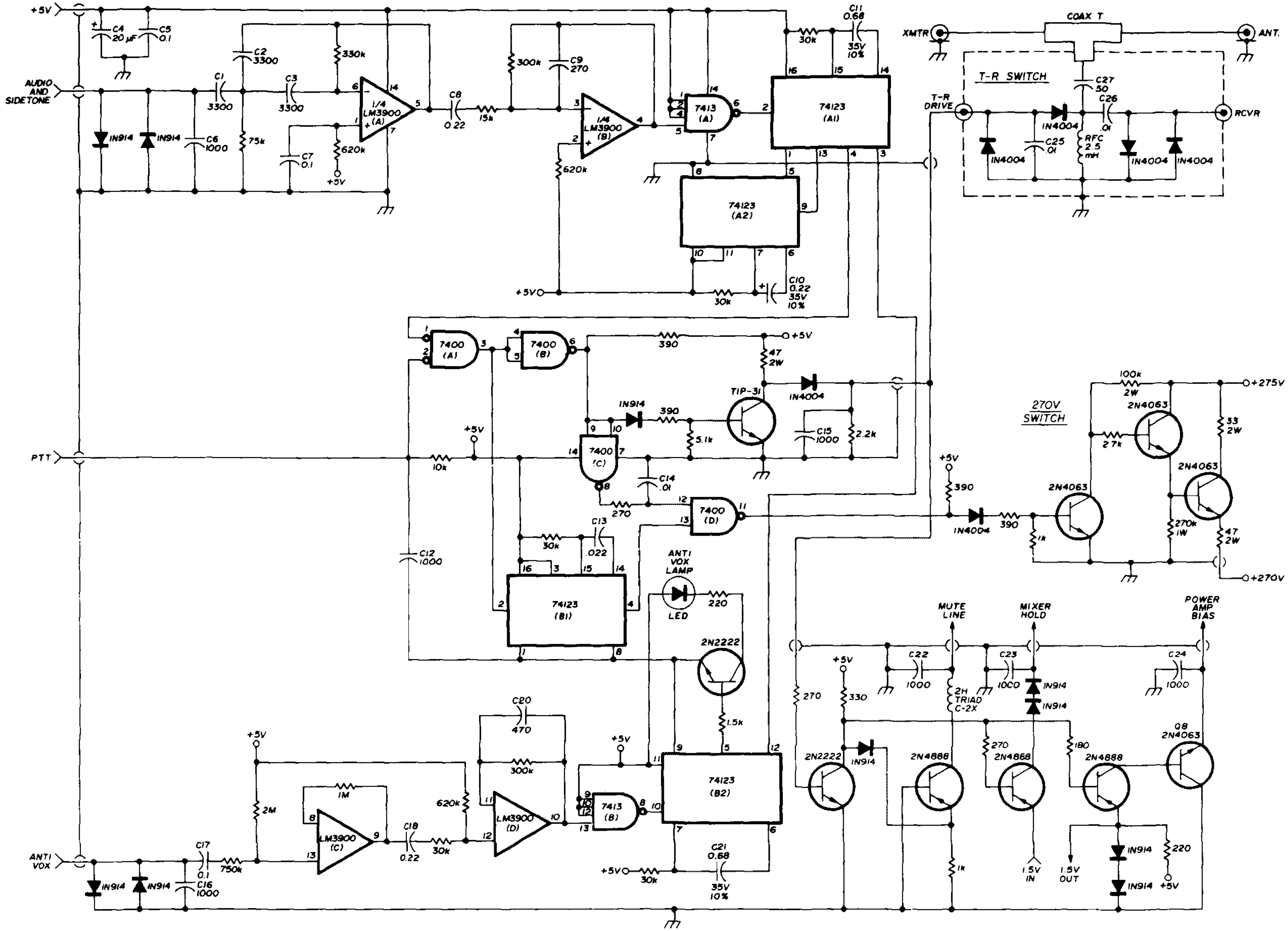



fig. 2. Schematic diagram of the control circuitry for the syllabic vox system. CR4-CR7, CR9, and CR10 are 1N4004 or equivalent diodes; all others are 1N914 or equivalent. The rf choke is a 2.5 mH, 200 mA transmitting type. All resistors are 1/2 watt, 10 per cent, unless otherwise specified.

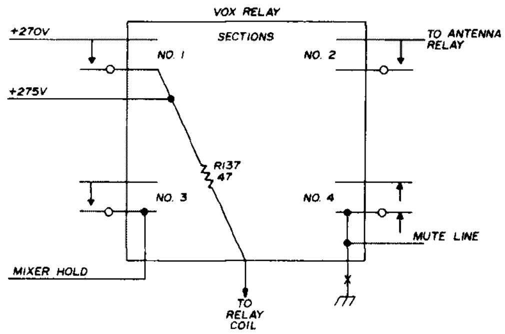



**fig. 3. View of the vox relay in the transmitter with the chassis upside down.**

6. One end of the 47-ohm, 2-watt resistor is connected to a relay coil terminal. The opposite end of the resistor is attached to the movable contact of section 1.

7. Remove the knobs from the vox and anti-vox controls. Also, remove the nuts so that the terminals of these potentiometers may be reached.

8. At the movable contact of the PTT switch (S10), connect the circuit board extension lead labelled "PTT", (**fig. 4A**).

9. At the wiper of vox gain control R74, connect the circuit board extension lead labelled "Audio and Sidetone".

10. At the wiper of anti-vox control, R85, connect the circuit board extension lead labelled "Anti-Vox".

11. Remove tubes V10, V14, and V12 since the high frequency oscillator of the receiver is ordinarily not used. This reduces the load on the power transformer.

12. Disconnect the plate lead to terminal 1 of V11.

13. Disconnect the lower end of R87 (470 ohms) from the PTT line. This resistor is near S8.

14. Disconnect the lead from the two parallel 2-watt resistors, R89 and R112. These two are also near S8, and alongside R87.

15. At S8, disconnect the lead between terminal 12 wafer D and ground. Also open the lead between terminals 12 of wafers D and C, (**fig. 4B**).

16. At terminal 12 of wafer D connect a circuit board extension lead labelled "+ 1.5 V Out".

17. At the bus wire connecting terminals 7, 8, 9 and 10, wafer D, connect a circuit board extension lead labelled "+ 1.5 V In".

18. Disconnect the lead to the antenna relay jack. Connect a circuit board extension lead labelled "Power Amplifier Bias" to the antenna relay jack.

19. At the vox relay, ground the several sheaths earlier mentioned. Use the ground lead already disconnected from the "Mute Contact".

20. Remove the cover of the rf section. The five screws are removed to reach the antenna relay contacts.

21. Short the antenna relay rf contacts with a short piece of wire (the rf side is near the pi network).

22. Disconnect the lead attached to the jack marked "Receive Antenna". This lead is now the rf output.

## 30L-1 modifications

The Collins 30L-1 linear amplifier requires a slight

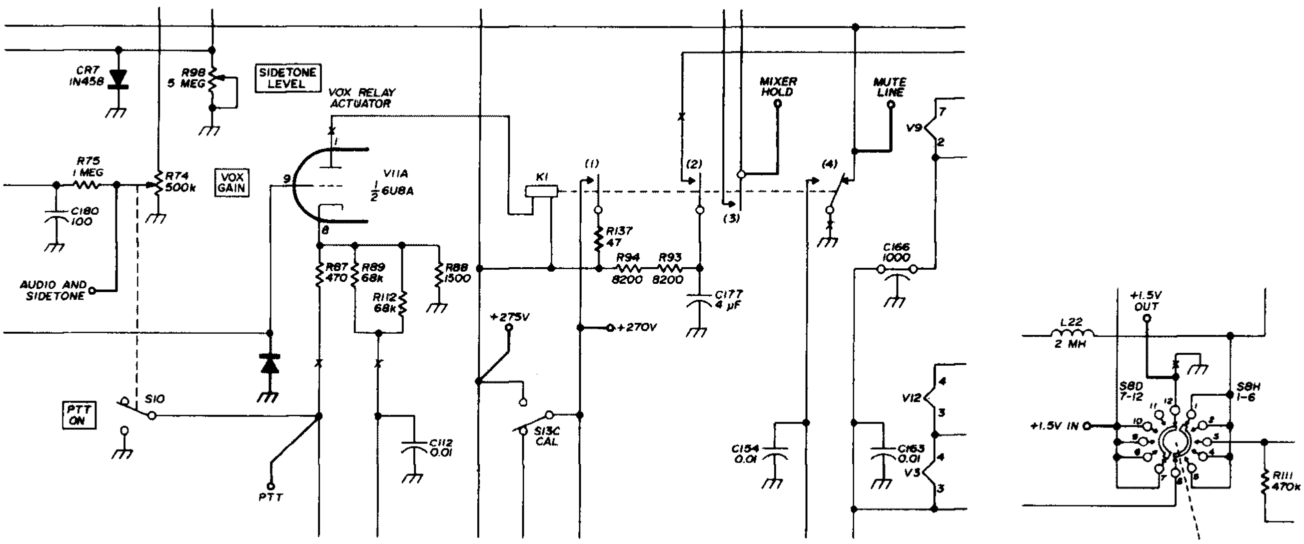



**fig. 4. Circuit modifications for the 32S-3 transmitter which will permit instantaneous vox operation.**

modification in the grid bias stage. Referring to **fig. 5**, the changes in the standard wiring diagram are indicated. The only components to be added to this area are as follows: 22k, 2-watt resistor; 1N4004 diode; and the 200 $\mu$F, 200-volt electrolytic capacitor. Be sure to observe the polarity of the diode connections. The revised circuit now uses a switching transistor to control cut-off during receive. The bias line is grounded through Q8 (2N4063) in accordance with transmitted voice syllables or code elements. The relay in the 30L-1 (K1) is now wired to be continuously energized; this relay is controlled with the main power switch. The large 200 $\mu$F capacitor is held in place with epoxy cement in the power compartment.

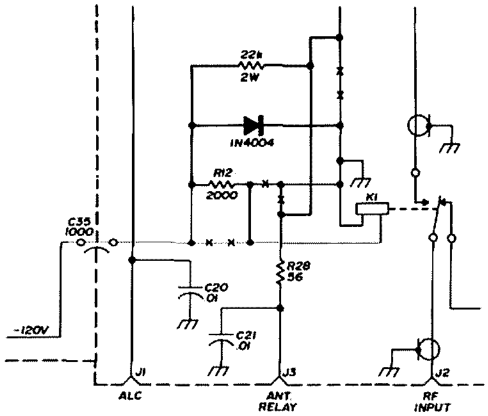



**fig. 5. The bias circuit in the Collins 30L-1 amplifier requires changes to interface with the transistor in the control circuitry. The 200 $\mu$F capacitor can be epoxied to the chassis.**

Some of the newer amplifiers have their grid bias lines wired in a similar fashion. The purpose of the modification is to maintain the tubes at cut-off bias during reception. The duty cycle during CW or ssb is short so the tubes operate at a much cooler level.

## results

For the past nine years I have been operating with equipment using no relays, in both the CW and ssb modes. The many advantages of high-speed, silent switching are numerous and add greatly to the pleasure of operating. The Collins conversion has added to that pleasure and has been most gratifying. My sincerest thanks are extended to W6RM, for without his guidance, the conversion would never have been accomplished.

## references

1. R. W. Hitchcock, W6RM, "Syllabic Vox System for the Drake Equipment," *ham radio*, August, 1976, page 24.
2. H. R. Hildreth, WØIP, "An Experimental All-Electonic Vox System for SSB," *QST*, March, 1968, page 36; "Instantaneous Voice Interruption," *QST*, October, 1968, page 40; "More on Instant Voice Interruption," *QST*, June, 1972, page 19.

**ham radio**

# hybrid-tee mixer for 10 GHz

How to use a hybrid-T waveguide junction to build a balanced mixer for 10-GHz receivers

The majority of amateur microwave receivers use front-ends which consist of a local oscillator coupled to a mixer diode. Radio-frequency amplifiers are available for the microwave frequencies but they are usually too expensive for amateur use. Receiver performance is therefore limited by the noise figure of the mixer, and to some extent by the following i-f amplifier, so any method which can improve mixer performance is well worth it.

At present two mixer types are prevalent among amateurs working at 10 GHz. The first is the self-oscillating mixer[1] (usually a Gunn device) which has the advantage of simplicity; the second is the mixer diode with a separate local oscillator. In the latter system optimum mixer performance can be obtained by adjusting local-oscillator injection and mixer impedance matching.

Some improvement in mixer performance can be achieved by using a balanced mixer. The design described in this article offers some improvement in mixer performance, and is simple to build.

## design

The hybrid-T waveguide junction shown in **fig. 1** has the following properties: power fed into port 3 will divide equally and in phase between ports 1 and 2; no power will appear at port 4. Power fed into port 4 will also divide equally between ports 1 and 2, but out of phase. Therefore, a balanced mixer may be constructed by placing mixer diodes at ports 1 and 2, injecting local oscillator power via port 3, and connecting the antenna to port 4. This arrangement has the advantage over the single-ended mixer in that the a-m noise sidebands associated with the local oscillator are cancelled out in the mixing process. In the single-ended mixer the local-oscillator noise sidebands are transferred to the i-f signal in the mixing process. System sensitivity is therefore limited by the quality of the local-oscillator signal. Further information on balanced mixers can be found in **references 2 and 3**.

The hybrid-T junction has useful properties as a power divider and isolator and may be constructed as a unit on its own. In commercial models it is usual

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to insert irises and matching stubs to compensate for the mismatch at the junction of the four ports. In the unit described here no attempt has been made to compensate for the inherent mismatch.

It is also necessary to match the mixer diodes to the waveguide. This is achieved by placing matching stubs (tuning screws) in front of the diode mounting. An improved match could be obtained by using a tapered diode mounting, however, the matching stub approach simplifies construction.

## construction

The mixer assembly is made from three pieces of waveguide (0.4 x 0.9 inch or 1 x 2.3cm internal dimensions) as shown in **fig. 2.*** The exact lengths are not critical but piece 1 should be at least 3¼ inches (8.3cm) long. Start by cutting the waveguide to length and neatly finishing off the ends. The slots in piece 1 should be scribed using the external dimensions of the waveguide for reference. The best way to remove the slots is to insert into the waveguide a piece of wood trimmed to the internal dimensions of

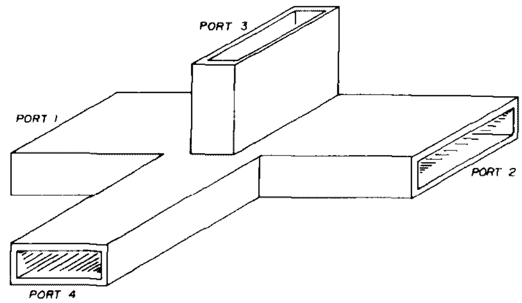


**fig. 1. The hybrid-T waveguide junction. Power fed into port 3 will divide equally and in phase between ports 1 and 2; no power will appear at port 4. Power fed into port 4 will also divide equally between ports 1 and 2, but out of phase. This property can be used to build a balanced mixer: the mixer diodes are installed at ports 1 and 2, local-oscillator power is injected at port 3, and the antenna is connected to port 4.**

adjacent waveguide wall because this will introduce discontinuities into the guide and permit solder to flow into the interior.

The slots should be filed out for a snug fit. Check that pieces 2 and 3 do not protrude into the inside of

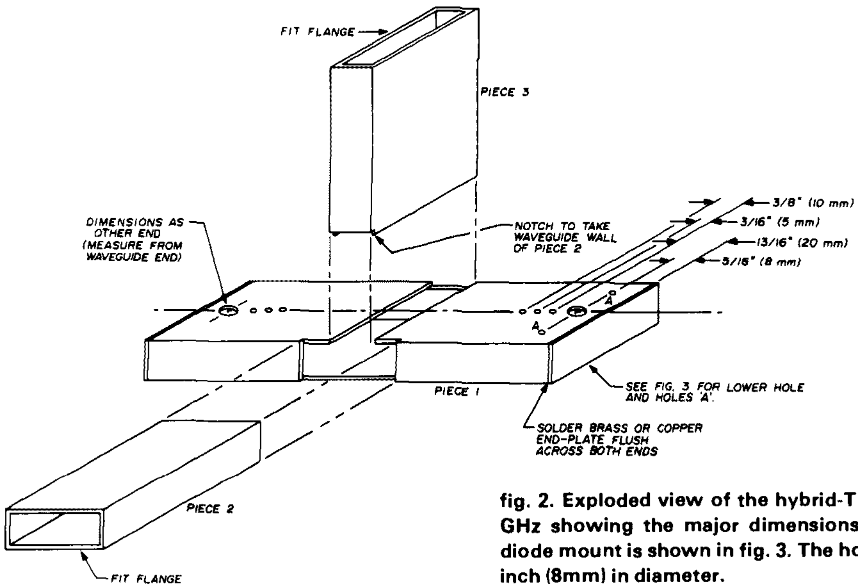


**fig. 2. Exploded view of the hybrid-T balanced mixer for 10 GHz showing the major dimensions. Construction of the diode mount is shown in fig. 3. The hole for the diode is 5/16 inch (8mm) in diameter.**

the guide. A neat cut will be obtained by using a small hacksaw and sawing on the inside of the scribed lines. It is important not to cut too deeply — the presence of sawdust in the metal cut will assist in judging the depth of the cut. Do not cut into the guide by looking into the open end of piece 1. When a good fit has been achieved, drill the matching screw holes and tap to suit the thread.*

The diode holes should be made by first drilling the smaller hole through both broad faces in one operation, this will ensure good alignment. The hole for the body of the diode can be entered as shown or from the opposite broad wall — this is a matter of choice.

*Following is a list of some of the common classifications used for 10-GHz waveguide. If in doubt check the internal dimensions — they should measure 0.4 x 0.9 inch (1x2.3cm).

| | |
|---|---|
| WG16 | United Kingdom |
| WR90 | United States (EIA) |
| RG-52/U | United States (JAN) |
| R100 | IEC |

*Any size screw thread of about 1/16" (1.5mm) diameter may be used.

The diode mounts are very simple in construction, but a word of caution: some diodes available on the surplus market are electrically sound but are out of specification mechanically. Therefore, use the dimensions given here only a s a guide and check with your favorite diodes for a good fit. The diode should be a firm sliding fit into the collar (**fig. 3**). The collar height should be such that the diode, when fitted in the waveguide, has its shoulder at the pin end touching the lower interior guide wall.

The insulation between the diode mount and the waveguide is ordinary cellophane tape (*e.g., Scotch* tape). The nylon retaining screws should be cut so they are flush with the inside wall of the waveguide when secured. When all drilling has been completed and all internal burrs are removed, soldering can be started. Pieces 1, 2, and 3 should be soldered first, followed by the waveguide flanges. A damp rag should be wrapped around the T junction to prevent

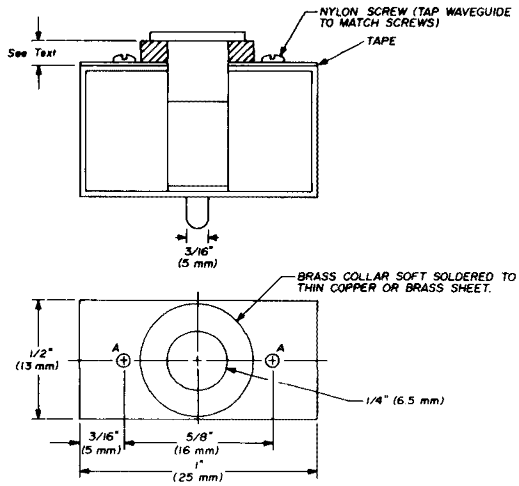


**fig. 3. Construction of the diode mount for the 10-GHz balanced mixer. The i-f connection is made to the diode through a solder lug placed under the head of one of the nylon screws (see fig. 4 for i-f and metering connections).**

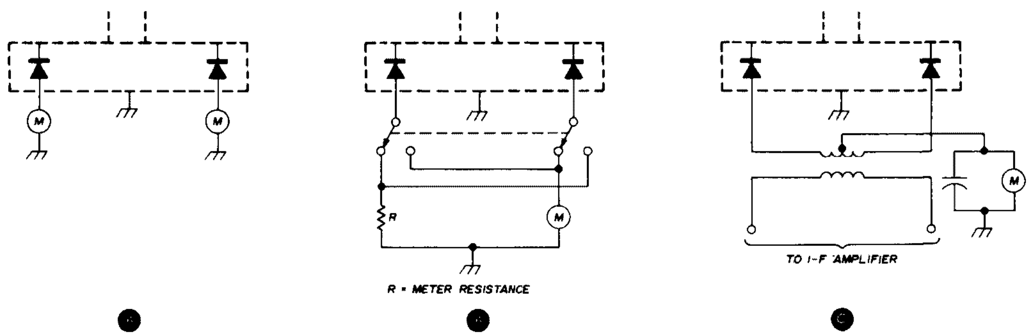


**fig. 4. Metering and i-f connections to the hybrid-T waveguide balanced mixer. When initially setting LO injection, two identical meters should be used as shown at A, although the metering circuit at B is acceptable. Connections to the i-f transformer are shown at C.**

the joint falling apart when soldering the flanges and end-pieces.

All soldering is best carried out with the work placed on a flat sheet of aluminum. This provides a firm base for the work, and avoids the possible embarrassment of soldering the waveguide to unwanted objects. A small gas jet should be used in preference to a soldering iron.

## adjustment

When connected to a local oscillator (5 -10 mW) the tuning screws should be adjusted for maximum mixer current. Ideally, this should be done with identical meters in series with each diode (**fig. 4A**), but if this is not possible the arrangement shown in **fig. 4B** can be used.

The tuning screws should be held in position with lock nuts. After the local-oscillator level has been adjusted for the recommended diode current, the metering circuit can be removed. A single meter for monitoring purposes can be placed in series with the i-f transformer center-tap (**fig. 4C**).

## conclusion

Performance will depend on the diodes you use (preferably a matched pair). No performance figures are given because a meaningful set of measurements is beyond my present resources; the main object has been to describe the construction of a simple balanced mixer for 10 GHz which will encourage other amateurs to copy, experiment, and hopefully improve the design.

## references


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2. F. E. Terman, *Electronic and Radio Engineering*, McGraw-Hill, New York, 1955, section 5-9.
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**ham radio**

# calculator-aided circuit analysis

Techniques to use your programmable, hand-held calculator to analyze the operation of your own circuits — before you even pick up your soldering iron

**Circuit analysis gives the** frequency response of components selected by circuit design. Design tells you what the circuit *should* do while analysis tells you what *actually* happens. Analysis can be done with a programmable pocket calculator on filters, matching networks, or other circuits with the same accuracy as big computers — and at far less cost. Programs and methods for the HP-25 programmable calculator are given here but there is enough data for conversion to other calculator types.

A programmable calculator allows accurate analysis of a circuit before it is built. Analysis takes a little time but it can literally save days otherwise spent in fiddling with marginal or unusable circuits. An advantage is that you get to know your circuits rather than just putting components together. It works with your own design or with published circuits.

The HP-25 can be purchased for less than $150 — about the average day's computer time charge — and the built-in functions are good for complex number handling. It is assumed you already know impedance and admittance, and rectangular and polar forms. All are needed to understand the methods. If these are unfamiliar, take time for a review before going on.

## basic ladder model

A *model* for circuit analysis is a block diagram made from *branches* connected by *nodes*. The branches are component combinations having two common connection points or nodes. A branch may have one or many components and its impedance or admittance value is described by a single expression. Any circuit may be modeled in this manner.

The *ladder* configuration is a shunt-series-shunt branch arrangement as shown in **fig. 1** for five branches. A ladder can be any odd-number branch arrangement. A model not in the ladder form can be transformed in subsections until the ladder is achieved. The simple ladder model allows only one signal source. This source is not considered a full branch.

**Fig. 1** shows all branches as impedances. Actual

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model values have shunt branches as admittances, series branches as impedances. This convention is used for the calculator solution program and the rationale may be seen by examining the model and identities plus the solution for output voltage.

Output voltage, $e_o$, will be a relative term and solely a function of branch values. Actual output voltage can be found by multiplying it by actual input current expressed in complex form. Most network responses are expressed in dB relative to a reference value so the current source can be unity without destroying validity.

There is one important fact about this model form: The admittance of the signal source *must* be a part of the input shunt branch. Excluding it will destroy the method of solving output voltage.*

Compare the terms in the final $e_o$ expression with the model diagram and identities. Each term represents the total admittance, $Y$, or impedance, $Z$, at a particular point in the model, looking towards the load. The progression of identities from $Y_{12}$ to $Z_{45}$ show this and the arrows of the diagram locate the position. By the time $Z_{45}$ is found, it will describe the *total* network impedance seen by the current source. Each successive $Y$ and $Z$ is both the sum of all preceding values and a term of the output voltage expression.

You can solve for both $e_o$ and total $Z_{in}$ by setting up the calculator to repeatedly request branch value entry, adding the entry to existing values, inverting the sum, multiply partial products by $e_o$ by the sum (accumulating), then returning for next branch entry. This follows the identities of **fig. 1** and is the main *loop* (repeated operations) or **Program 1**.

A beginner in step-programming may be a little confused when confronted by only a program. The flow chart in **fig. 2** represents the ladder network calculation of **Program 1**.

## ladder network flow chart

Flow charts are to programming what schematics are to circuits. Each chart symbol represents a component of the program. Lines with arrows indicate progression from one operation to the next. Rectangles with square sides are general operations, rounded-side boxes are program stops to display data, and slanted-side boxes are data entry stops. Diamonds represent decision points in the flow; program flow can change depending on what decision is made with particular data.

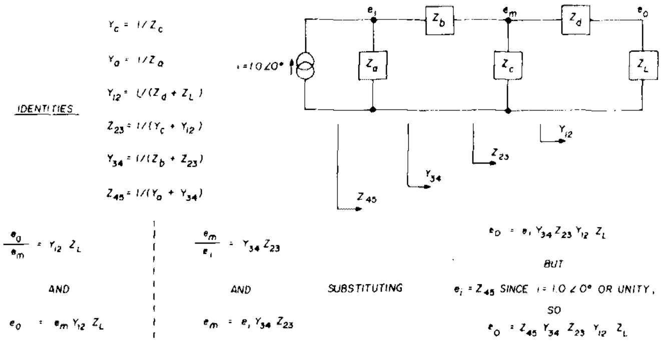



**fig. 1. Basic ladder network which can be quickly and easily analyzed with a hand-held, programmable calculator.**

The loop is at upper left. Branch value entry uses the rectangular form while inversion and multiplication is done in polar form. Partial products of $e_o$ accumulate in memory while successive $Y$ and $Z$ remains in the stack.† The loop must stop after the last entry calculation so a counter is used for control.

The counter is just a memory register. Initialization at program start sets the counter to the number of branches in the model. After each loop is complete, the counter is decremented (reduced by one) and its value tested. A positive, non-zero value causes the program to jump back to loop start. Loop iteration stops when the counter is zero and program flow changes to the lower-right group.

The program is designed to give output voltage in dB relative to a reference, $e_{ref}$. Since the output voltage is unknown, reference to 0 dB is also unknown. The reference is automatically set if the first solution has the memory register for $e_{ref}$ preset to zero. In this case, $e_o$ is compared to $e_i$ (input voltage = total

Computer Aided Design, or CAD as it's known in industry, has been widely used by engineers and researchers to design and analyze electronic circuits. With the availability of low-cost, hand-held, programmable calculators, many of the benefits of this powerful analytical tool are now available to amateurs and experimenters.

In 1970 the author had to quickly find some time-delay equalizing networks for an avionics system's detector bank. Time was short and some components were in stock that *might* work. But would they? A fellow engineer told Anderson about the company's computer program and helped him check the circuit out on the terminal. The computer program had time-response capability and a half-hour of printout said the circuit would work fine. The circuit was built in the lab and checked out — the scope display was *identical* to the computer printout. The total elapsed time from assignment to results was less than five hours. The author has been a strong advocate of CAD ever since.

**Editor**

*Exclusion of signal-source admittance does not affect input impedance.

†An algebraic-type calculator requires different storages so the program must be written from the original formulas of **fig. 1**.

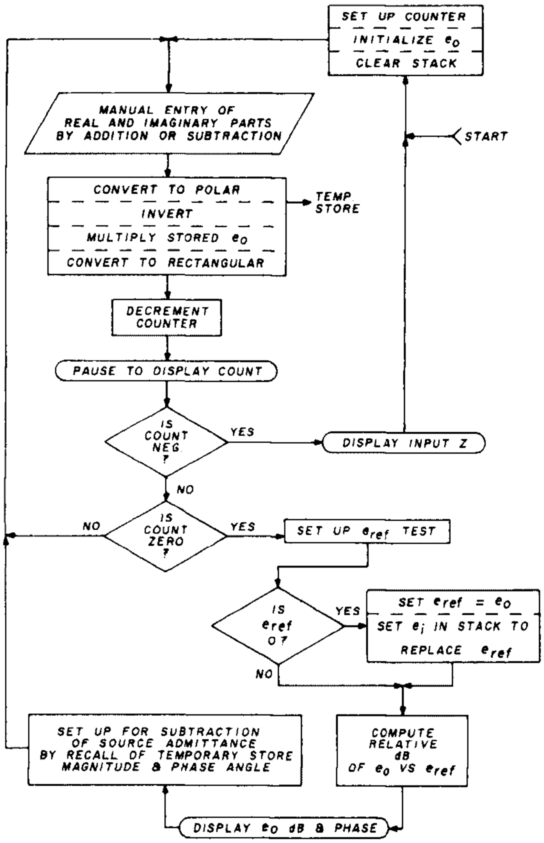



**fig. 2. Flow chart for Program 1, ladder network analysis.**

input impedance) and the first $e_o$ display represents insertion loss/gain. All other $e_o$ displays compare to $e_{ref}$ unless the memory register is reset. An analysis usually requires one frequency at which relative output is 0 dB so this frequency should be solved first.

The lower-left block is a special operation. This sets up conditions for subtracting source impedance from total network admittance to find input impedance of only the network. This is very useful for filters and other circuits that require input impedance data. Each successive $Y$ or $Z$ calculated in the loop is also placed in temporary memory storage. In this part of the program the temporary storage contains total input admittance so this is recalled to the stack and flow jumps to loop start. At loop finish the counter will be $-1$ so the flow is directed to input impedance display.

## program listing and details

The memory register block of **Program 1** should be noted. Only three registers, R4, R5, and R6 must be preset before any run. Register R4 contains the constant 20 used in computing voltage gain in dB; holding constants in memory saves program steps. R5 is initially zero to establish the 0 dB reference voltage for subsequent frequency solutions. R6 is set to the number of branches (do not count the current source as a branch).

Registers 0 and 1 accumulate the output voltage. Output phase angle is proportional to zero-degree input *current*. Input *voltage* phase angle is a function of total input impedance phase angle and must be calculated outside of this program. Successive $Y$ and $Z$ from the loop are stored in R2 and R3.

**Steps 01** through **06** initialize the program run (upper right-hand corner of **fig. 2**). Register **R0** is set to one so the partial products of $e_o$ can be accumulated by register arithmetic. Register arithmetic allows any memory register to be added, subtracted, multiplied, or divided by the X register content without altering the X register.

The loop is contained in **steps 07 - 21. Steps 07** and **08** do nothing on the first pass through the loop since both X and Y stack registers are zero. These steps are required on return jumps and will then contain the data indicated.

Branch value entry must be done manually by addition or subtraction, depending on sign. This is required to fit the limited number of program steps. Order of entry is always real part first, imaginary part second. Rectangular form is used here for both program and value-precalculation convenience.

Output voltage partial products are accumulated at **steps 16** and **20** using register arithmetic. Polar form is more convenient here. Inversion of the polar form

| STEP | KEY | X | Y | Z | T | REMARKS | |
|---|---|---|---|---|---|---|---|
| 01 | RCL 6 | NB | | | | INITIALIZE | |
| 02 | STO 7 | NB | | | | COUNTER | |
| 03 | 1 | 1 | NB | | | | |
| 04 | STO 0 | 1 | NB | | | INIT. $e_o$ MAGN. | |
| 05 | f STK | 0 | 0 | 0 | 0 | CLEAR STACK | |
| 06 | STO 1 | 0 | 0 | 0 | 0 | INIT. $e_o$ PHA | |
| 07 | R↓ | (MAG) | (PHA) | | | LOOP START | |
| 08 | f →R | REAL | IMAG | | | RECT. CONVERT | |
| 09 | R/S | REAL | IMAG | | | INPUT REAL | ◀ENTER BY |
| 10 | x⇄y | IMAG | REAL | | | | MANUAL ADD |
| 11 | R/S | IMAG | REAL | | | INPUT IMAG | ◀OR SUBTRACT |
| 12 | x⇄y | REAL | IMAG | | | | |
| 13 | g →P | MAG | PHA | | | POLAR CONVERT | |
| 14 | g 1/x | 1/MAG | PHA | | | BEGIN INVERT | |
| 15 | STO 2 | 1/MAG | PHA | | | TEMP. STORE | |
| 16 | STO × 0 | 1/MAG | PHA | | | ACC. $e_o$ MAGN. | |
| 17 | x⇄y | PHA | 1/MAG | | | | |
| 18 | STO 3 | PHA | 1/MAG | | | TEMP. STORE | |
| 19 | CHS | −PHA | 1/MAG | | | FINISH INVERT | |
| 20 | STO + 1 | −PHA | 1/MAG | | | ACC. $e_o$ PHA | |
| 21 | x⇄y | 1/MAG | −PHA | | | | |
| 22 | RCL 7 | N | 1/MAG | −PHA | | DECREMENT COUNTER | |
| 23 | 1 | 1 | N | 1/MAG | −PHA | | |
| 24 | − | N−1 | 1/MAG | −PHA | | | |
| 25 | STO 7 | N−1 | 1/MAG | −PHA | | | |
| 26 | f PAUSE | N−1 | 1/MAG | −PHA | | (OPTIONAL) | |
| 27 | g x<0 | N−1 | 1/MAG | −PHA | | | |
| 28 | GTO 49 | N−1 | 1/MAG | −PHA | | JMP Z DISPLAY | |
| 29 | g x≠0 | N−1 | 1/MAG | −PHA | | | |
| 30 | GTO 07 | N−1 | 1/MAG | −PHA | | JMP LOOP START | |
| 31 | RCL 0 | $e_o$ | 0 | 1/MAG | −PHA | | |
| 32 | RCL 5 | $e_{ref}$ | $e_o$ | 0 | 1/MAG | | |
| 33 | g x≠0 | $e_{ref}$ | $e_o$ | 0 | 1/MAG | | |
| 34 | GTO 38 | $e_{ref}$ | $e_o$ | 0 | | JMP DB OUTPUT | |
| 35 | R↓ | $e_o$ | | | | | |
| 36 | STO 5 | $e_o$ | | | | SET NEW REFER. | |
| 37 | RCL 2 | $e_i$ | $e_o$ | | | | |
| 38 | ÷ | RATIO | | | | | |
| 39 | f log | LOG | | | | | |
| 40 | RCL 4 | 20 | LOG | | | | |
| 41 | × | DB | | | | | |
| 42 | R/S | DB | | | | DISPLAY $e_o$ MAG | |
| 43 | RCL 1 | PHA | DB | | | | |
| 44 | R/S | PHA | DB | | | DISPLAY $e_o$ PHA | |
| 45 | RCL 3 | Y PHA | | | | SETUP TO FIND | |
| 46 | RCL 2 | Y MAG | Y PHA | | | NETWORK INPUT | |
| 47 | g 1/x | Y MAG | Y PHA | | | IMPEDANCE | |
| 48 | GTO 08 | Y MAG | Y PHA | | | JMP LOOP CONV. | |
| 49 | R↓ | Z MAG | Z PHA | | | DISPLAY Z IN | RESTART BY R/S |

MEMORY

| | |
|---|---|
| R0 | $e_o$ MAGNITUDE |
| R1 | $e_o$ PHASE ANGLE |
| R2 | MAGNITUDE (TEMPORARY) |
| R3 | PHASE ANGLE (TEMPORARY) |
| R4 | 20 (CONSTANT) |
| R5 | $e_{ref}$ (REFERENCE) |
| R6 | NUMBER OF BRANCHES (CONSTANT) |
| R7 | N COUNTER |

**Program 1. Ladder Network Calculation.**

is at **steps 14** and **19**. The intermix of steps may be confusing at first glance but is required to reduce the program size.

The pause at **step 26** is optional but useful when becoming acquainted with the program. The one-second display will show the branch number to be entered *next*. This assumes the first shunt branch is 1, the next series branch is **2**, and so on. A zero display indicates that entries are complete and $e_o$ display is next.

The first decision point is at **steps 27** and **28** on the new state of the counter contained in X. Program flow can be stated as,"If the contents of X is less than zero (negative), then go to **step 49**; otherwise continue with statement in **step 29**." The next decision point follows directly and is interpreted, "If the contents of X is *not* zero, then go to **step 07**; otherwise continue with **step 31**."

Many functions on the HP-25 require the *g* or *f* key prefix so make certain you use the correct one when entering the program steps. Strange results could happen from **steps 27** and **29** if the *f* key was mistakenly used.

**Steps 31** and **32** set up the $e_{ref}$ test actually performed in **steps 33** and **34**. The 0 shown in Y and Z stack registers is the counter state; the program would not have reached **step 31** unless the counter was zero. Reference test is done on polar form magnitudes only. The content of R2 is now equal to total $Z_{in}$ or equal to $e_i$. **Step 37** and the register roll-down of **step 35** are equivalent to replacing $e_{ref}$ by $e_i$.

**Steps 38 - 41** compute voltage ratio in dB, displayed at **42** stop. Any register may now be examined manually since the following steps order the stack contents by recall. This is also true after the phase angle display stop at **step 44**.

**Steps 45 - 48** set up for source admittance subtraction. Phase angle in register R3 is in admittance form since the storage instruction of **step 18** came before the **step 19** sign change for inversion completion. R2 is an impedance since it was stored after inversion; this was a requirement of recall at **step 37** to reduce size. **Step 47** completes the magnitude inversion so the jump to **step 08** has X and Y in admittance form. Source admittance is subtracted within the loop and the resulting impedance magnitude is displayed at **step 49**. Exchanging X and Y manually will display impedance phase angle. Depressing the *R/S* key will return to **step 01** restarting for the next run.

Precalculation of all branch values is required before a solution is done. All frequency-sensitive branches must be precalculated at each solution frequency. If either real or imaginary part of the branch is zero, the particular entry may be skipped. It is important to keep entry order of real-part-first in mind; solution is invalid if these are mixed.

**table 1. Keyboard operations and displays using HP-25 Program 1 to analyze the R-C lowpass filter shown in fig. 3.**

| keyboard | display | |
|---|---|---|
| | 0.0000 | (first stop after restart) |
| enter load G | 1.     -05 | (EEX, CHS, 5) |
| add | 1.0000000-05 | |
| R/S | 0.0000 | |
| enter 0.005 μF B | 0.000062832 | |
| add | 0.0001 | |
| R/S | 4.0000 | (pause display for 1 second) |
| | 2470.4410 | |
| enter 10k res. | 1.     04 | (EEX, 4) |
| add | 12470.4410 | |
| R/S | - 15522.2751 | |
| (skip entry, X = 0) | | |
| R/S | 3.0000 | (pause display for 1 second) |
| | 3.1455036-05 | |
| (skip entry, G = 0) | | |
| R/S | 3.9152883-05 | |
| enter 0.05 μF B | 0.000628319 | |
| add | 0.0007 | |
| R/S | 2.0000 | (pause display for 1 second) |
| | 70.4467 | |
| enter 1k res. | 1.     03 | (EEX, 3) |
| add | 1070.4467 | |
| R/S | - 1494.8706 | |
| (skip entry, X = 0) | | |
| R/S | 1.0000 | (pause display for 1 second) |
| | 0.0003 | |
| enter source G | 0.01 | |
| add | 0.0103 | |
| R/S | 0.0004 | |
| (skip entry, B = 0) | | |
| R/S | 0.0000 | (pause display for 1 second) |
| | - 3.2067 | $e_o$ magnitude in dB |
| R/S | - 65.0970 | $e_o$ phase angle in degrees |
| R/S | 0.0103 | total input conductance |
| enter source G | 0.01 | |
| subtract | 0.0003 | |
| R/S | 0.0004 | total input susceptance |
| (skip entry, B = 0) | | |
| R/S | - 1.0000 | (pause display for 1 second) |
| | 1838.6121 | filter input $Z$ magnitude |
| x ↔ y | - 54.3943 | filter input $Z$ phase angle |
| RCL 5 | 90.0090 | $e_{ref}$ |

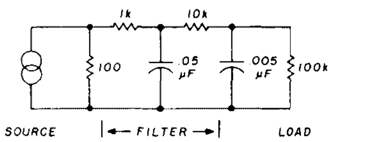
## example using program 1

The two-section R-C lowpass filter of **fig. 3** is a simple example. Capacitors are assumed lossless. There are only five branches in the model. The first branch has an admittance of $0.01 + j0$ *mho* due to the source resistance of **100** ohms. Load admittance of **10** micromho is constant in branch **5** but the susceptance of the **0.005** μF is in parallel. Branches **2** and **4** are purely resistive while branch **3** (0.05 μF) is purely susceptive. To reiterate, all shunt branches are entered as admittances while series branches enter as impedances.

**Fig. 3** shows a tabulation of capacitor susceptance

as well as solved output voltage and filter input impedance. Susceptance is zero at dc, the reference frequency, so the reference solution required only resistive or conductive values entries. Insertion loss will display negative but some circuits have voltage gain; a voltage gain displays positive at the reference frequency.

Memory register 6 is set at 5 for the number of branches and R5 is set at zero for the reference run at dc. A tabulation of keyboard operations and resulting displays are shown in **table 1**. Display is set fixed at 4-decimals and the run is at 2 kHz.




| frequency (kHz) | susceptance (mho) 0.05 μF | 0.005 μF |
|---|---|---|
| 1 | 0.000314159 | 0.000031416 |
| 2 | 0.000628319 | 0.000062832 |
| 3 | 0.000942478 | 0.000094248 |
| 4 | 0.001256637 | 0.000125664 |
| 5 | 0.001570796 | 0.000157080 |
| 10 | 0.003141593 | 0.000314159 |
| 20 | 0.006283186 | 0.000628319 |

| frequency (kHz) | output voltage dB | phase angle | input | filter impedance |
|---|---|---|---|---|
| 0* | 0.0 | 0.0000° | 111000 | 0.0000° |
| 1 | −0.9499 | −36.0202° | 3162.84 | −68.8897° |
| 2 | −3.2067 | −65.0970° | 1838.61 | −54.3943° |
| 3 | −5.8972 | −86.4955° | 1456.26 | −43.9354° |
| 4 | −8.5755 | −102.1197° | 1288.50 | −36.5079° |
| 5 | −11.0862 | −113.7452° | 1198.50 | −31.0653° |
| 10 | −20.7912 | −143.2080° | 1057.05 | −17.3493° |
| 20 | −32.1027 | −160.9919° | 1014.94 | −8.9974° |

*Reference frequency, insertion loss = 0.09065 dB

**fig. 3. Susceptance, output voltage, and filter input impedance for an R-C lowpass filter, calculated with a programmable, hand-held calculator.**

The tabulation looks formidable but operation takes less time than reading the listing. The RCL 5 at the end is not required; it was included only to show the internal $e_{ref}$ value in memory.

For simple circuits such as this one, 5 significant digits of entry value should be sufficient. At least 6 digits should be used for larger models or those having reactance and susceptance in each branch. Most big computer analysis programs have only 7 significant digits for all calculation while the HP-25 uses ten — the program can actually be more accurate than computer results!*

Analysis of L-C networks, especially filters, can be tiring when R or G is absent or a constant value. **Program 2** was designed for filter analysis when the source and load resistance is equal and the shunt branch conductance is always the same value. Top-coupled bandpass filters can be analyzed by this program. In this case the shunt conductance is calculated by $B_L/Q$ at center frequency and $Q$ is the mean value of parallel inductor and capacitor.

## modified ladder network program

**Program 2** is a rewritten version of **Program 1** and requires only the imaginary part of branch value be entered. Series branches are assumed nearly lossless; common shunt conductance is also added to the series branches by the program so solutions are valid only when series reactance is high compared to parallel reactance. It does serve to show how the basic program can be changed to achieve the same results.

Initialization, loop, and output voltage calculation in dB are the same as **Program 1** with a few exceptions. There is no temporary storage since this is occupied by constants $1/R_L$ and $G_s$, the common shunt conductance. This leaves no room for finding $Z_{in}$ by the program or automatically setting an $e_{ref}$ value. Each of these can be set manually on the first run. The loop of **Program 1** starts and ends in polar form. **Program 2's** loop starts and ends in rectangular form so that shunt conductance and end conductance may be added by the program.

A notable exception is that the number of branches stored in R6 is one less than the model. This is a requirement of steering program flow at **steps 29** and **30** so that end conductance can also be added via **steps 32 - 34**. Load-end conductance is added by the GTO32 at **step 07** while source-end conductance addition results from decision steering.

A negative count at loop ending steers the program to output voltage calculation. The initial value of $e_{ref}$ should be some number like one to prevent an error stop at **step 38** from division by zero. The first $e_o$ display will be meaningless.

Total input impedance has been carried in the stack in rectangular form. This is converted to polar form after output voltage phase angle display. Manual register operations should be avoided until the stop at **step 48** since stack data could disappear.

*The question of accuracy sometimes comes up when calculator results are compared to computer printouts. Large computers don't automatically imply great accuracy — only when *long precision* (10 digits) or *double precision* (14 digits) are used does this happen. A recent computer run on an audio lowpass filter for an ssb rig indicated a *dc* insertion loss of 0.0002 dB with a −0.3723° phase angle! Since inductor $Q$ was not modeled in this program, insertion loss should be 0 dB. The computer program's phase angle can only be explained by the way the program was written and by the 7-digit precision.

| STEP | KEY | X | Y | Z | T | REMARKS |
|---|---|---|---|---|---|---|
| 01 | RCL 6 | NB | | | | INITIALIZE |
| 02 | STO 7 | NB | | | | COUNTER |
| 03 | 1 | 1 | NB | | | |
| 04 | STO 0 | 1 | NB | | | INIT. $e_o$ MAG. |
| 05 | f STK | 0 | 0 | 0 | 0 | CLEAR STACK |
| 06 | STO 1 | 0 | 0 | 0 | 0 | INIT. $e_o$ PHA |
| 07 | GTO 32 | 0 | 0 | 0 | 0 | JMP TO GL ADD |
| 08 | R↓ | REAL | IMAG | | | LOOP START |
| 09 | RCL 3 | GS | REAL | IMAG | | |
| 10 | + | REAL' | IMAG | | | ADD GS |
| 11 | x⇄y | IMAG | REAL' | | | |
| 12 | R/S | IMAG | REAL | | | INPUT IM BY Σ |
| 13 | x⇄y | REAL | IMAG | | | |
| 14 | g →P | MAG | PHA | | | POLAR CONVERT |
| 15 | g 1/x | 1/MAG | PHA | | | BEGIN INVERT |
| 16 | STO × 0 | 1/MAG | PHA | | | ACC. $e_o$ MAGN. |
| 17 | x⇄y | PHA | 1/MAG | | | |
| 18 | CHS | -PHA | 1/MAG | | | FINISH INVERT |
| 19 | STO − 1 | -PHA | 1/MAG | | | ACC. $e_o$ PHA |
| 20 | x⇄y | 1/MAG | -PHA | | | |
| 21 | f →R | REAL | IMAG | | | RECT. CONVERT |
| 22 | RCL 7 | N | REAL | IMAG | | DECREMENT |
| 23 | 1 | 1 | N | REAL | IMAG | COUNTER |
| 24 | − | N-1 | REAL | IMAG | | |
| 25 | STO 7 | N-1 | REAL | IMAG | | |
| 26 | f PAUSE | N-1 | REAL | IMAG | | (OPTIONAL) |
| 27 | g x<0 | N-1 | REAL | IMAG | | |
| 28 | GTO 35 | N-1 | REAL | IMAG | | JMP OUTP.CALC. |
| 29 | g x≠0 | N-1 | REAL | IMAG | | |
| 30 | GTO 08 | N-1 | REAL | IMAG | | JMP LOOP START |
| 31 | R↓ | REAL | IMAG | | | |
| 32 | RCL 2 | GL | REAL | IMAG | | |
| 33 | + | REAL' | IMAG | | | ADD END GL |
| 34 | GTO 09 | REAL' | IMAG | | | JMP TO GS ADD |
| 35 | R↓ | REAL | IMAG | | | BEG.OUTP.CALC. |
| 36 | RCL 0 | $e_o$ | REAL | IMAG | | |
| 37 | RCL 5 | $e_{ref}$ | $e_o$ | REAL | IMAG | |
| 38 | ÷ | RATIO | REAL | IMAG | | |
| 39 | f log | LOG | REAL | IMAG | | |
| 40 | RCL 4 | 20 | LOG | REAL | IMAG | |
| 41 | × | DB | REAL | IMAG | | |
| 42 | R/S | DB | REAL | IMAG | | DISPLAY $e_o$MAG |
| 43 | R↓ | REAL | IMAG | | | |
| 44 | RCL 1 | PHA | REAL | IMAG | | |
| 45 | R/S | PHA | REAL | IMAG | | DISPLAY $e_o$PHA |
| 46 | R↓ | REAL | IMAG | | | |
| 47 | g →P | MAG | PHA | | | |
| 48 | R/S | MAG | PHA | | | DISPLAY Z MAG |
| 49 | x⇄y | PHA | MAG | | | DISPLAY Z PHA |

ENTER IMAGINARY PART BY MANUAL ADD OR SUBTRACT

RESTART BY R/S

MEMORY

| | |
|---|---|
| R0 | $e_o$ MAGNITUDE |
| R1 | $e_o$ PHASE ANGLE |
| R2 | 1/RL (GL) END CONDUCTANCE |
| R3 | GS COMMON SHUNT CONDUCTANCE |
| R4 | 20 (CONSTANT) |
| R5 | $e_{ref}$ (REFERENCE) |
| R6 | NUMBER OF BRANCHES MINUS ONE |
| R7 | N COUNTER |

**Program 2. Modified Ladder Network Calculation.**

Network input $Z$ must be calculated manually in this program.

Insertion loss at the reference frequency can be found as follows: Store total $Z$ magnitude in register R7 by STO 7 after the stop at step **48**. Continue to step **49** and record total $Z$ phase. Depress RCL **0**, STO 5 to set $e_{ref}$ in memory, then RCL 7 and DIVIDE to find $e_o/e_i$ ratio. Using R7 at this point will not disturb the program since a restart will initialize the counter, destroying any previous data there.

**Program 2** is convenient for branch entry but inconvenient for reference and network input impedance. Tradeoffs must be made due to program size. The last steps of both programs can be altered to suit the reader's application.

## precalculation techniques

Branch values at each solution frequency must be known before applying either **Program 1** or **Program 2**. This can be done by another calculator program and tabulation of values. Precalculation should begin at the last branch and work forward. Tabulation will then have the same order of entry as either solution program.

**Fig. 4** shows a simple bandpass filter to be analyzed. The model will have only three branches. Source and load resistances are equal and capacitor $Q$ is assumed much higher than inductor $Q$. Equivalent shunt conductance due to $Q$ is calculated from inductive susceptance only.

**Program 3** is an example of precalculation for fig. **4**. Frequency is the only variable and all component values are stored in memory. This lets the calculator do all the work of sign and decimal-point placement for the least error.

Register arithmetic is used extensively and programming is *linear* (no loops or jumps). The memory block shows two values for registers R0 through R5, excluding R2. The first value is the preset constant while the second is frequency modified. Note that inductance and C2 is preset inverted, C2 is also signed negative. Since calculation will require this anyway, it saves program steps to store constants in this form.

**Steps 02** through **08** change the memory by register arithmetic and also store the entered frequency. **Steps 33-39** restore memory contents to original value. Register arithmetic is quite useful for repetitive operations like this and saves program steps. A little study will show how it is done.

Inductive susceptance is shown with a negative sign. A negative $B_L$ is a positive quantity required for equivalent conductance due to $Q$. The subtraction at steps **17** and **31** properly sets the total susceptance sign. The negative $B_L$ is a subtle reminder to remember the basics of admittance.

Computation is straightforward with display stops in order from load to generator, real part first, imaginary part second. Shunt branches are in admittance, series in impedance. **Steps 19** and **20** are strictly

| STEP | KEY | X | Y | Z | T | REMARKS |
|---|---|---|---|---|---|---|
| 01 | R/S | FREQ | | | | ENTER |
| 02 | STO × 0 | FREQ | | | | RADIAN FREQ. |
| 03 | STO 7 | FREQ | | | | STORE TEMP. |
| 04 | RCL 0 | ω | FREQ | | | SET RAD.FREQ |
| 05 | STO ÷ 1 | ω | FREQ | | | FIND -B(L) |
| 06 | STO × 3 | ω | FREQ | | | FIND B(C1) |
| 07 | STO ÷ 4 | ω | FREQ | | | FIND X(C2) |
| 08 | STO × 5 | ω | FREQ | | | FIND B(C3) |
| 09 | RCL 1 | -B(L) | | | | BEGIN LOAD G |
| 10 | RCL 2 | QL | -B(L) | | | |
| 11 | ÷ | G(QL) | | | | |
| 12 | RCL 6 | G(LOAD) | G(QL) | | | |
| 13 | + | GL | | | | END LOAD G |
| 14 | R/S | GL | | | | |
| 15 | RCL 3 | B(C1) | | | | |
| 16 | RCL 1 | -B(L) | B(C1) | | | |
| 17 | − | BL | | | | |
| 18 | R/S | BL | | | | LOAD B |
| 19 | CLX | 0 | | | | |
| 20 | R/S | 0 | | | | SERIES R |
| 21 | RCL 4 | X(C2) | | | | |
| 22 | R/S | X(C2) | | | | SERIES X |
| 23 | RCL 1 | -B(L) | | | | BEGIN GENR. G |
| 24 | RCL 2 | QL | -B(L) | | | |
| 25 | ÷ | G(QL) | | | | |
| 26 | RCL 6 | G(LOAD) | G(QL) | | | |
| 27 | + | G(GEN) | | | | END GENR. G |
| 28 | R/S | G(GEN) | | | | GEN G |
| 29 | RCL 5 | B(C3) | | | | |
| 30 | RCL 1 | -B(L) | | | | |
| 31 | − | B(GEN) | | | | |
| 32 | R/S | B(GEN) | | | | GEN B |
| 33 | RCL 0 | ω | | | | |
| 34 | STO × 1 | ω | | | | MEMORY RESET |
| 35 | STO ÷ 3 | ω | | | | MEMORY RESET |
| 36 | STO × 4 | ω | | | | MEMORY RESET |
| 37 | STO ÷ 5 | ω | | | | MEMORY RESET |
| 38 | RCL 7 | FREQ | | | | |
| 39 | STO ÷ 0 | | | | | 2PI RESTORE |
| 40 | GTO 01 | | | | | JMP TO START |
| 41 | | | | | | |
| 42 | | | | | | |
| 43 | | | | | | |
| 44 | | | | | | |
| 45 | | | | | | |
| 46 | | | | | | |
| 47 | | | | | | |
| 48 | | | | | | |
| 49 | | | | | | |

MEMORY

| | |
|---|---|
| R0 | 2 PI / 2PI F (ω) |
| R1 | 1/L / -B(L) |
| R2 | QL |
| R3 | C1 / B(C1) |
| R4 | -1/C2 / X(C2) |
| R5 | C3 / B(C3) |
| R6 | 1/RL (GL) END CONDUCTANCE |
| R7 | FREQ (TEMPORARY) |

**Program 3. Precalculation Example.**

reminders that the real part of the series branch is zero; these can be deleted.

**Steps 23-28** are duplicates of **09-14** since both inductors are the same value. The program can be written to hold end conductance in the stack after step **14**, then bring it back for display. A better alternative is to let register R2 hold the constant $1/(L\ Q_I)$ and add two register arithmetic steps. This would delete four steps from the program shown. It is always better to hold constants in memory so the program needs fewer steps to compute the desired value.

The original frequency entry will be displayed when the run is finished and jumps back to stop at **step 01**. This is a good reminder for tabulation. If you hit the **R/S** key at **step 01** without entering a new frequency, all displays will repeat the previous run.

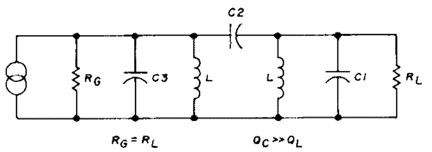



**fig. 4. Bandpass filter for the precalculation example discussed in the text.**

Run through all displays when this occurs so that memory is restored. A mid-program change requires reloading memory.

Many variations of the precalculation program are possible. They are written to fit the particular circuit to be modeled. Adopting the style of **Program 3** has these advantages:

1. Only one entry for frequency, all other stops are displays.
2. Component storage in memory with inversions and signs as necessary will reduce program steps.
3. Register arithmetic forms reactance and susceptance directly in memory with the fewest steps.
4. Calculation is simple and direct.

More than one program may be required. It is suggested that the load-end values are all calculated first. This fits the back-to-front ordering of the solution programs.

Scaling to megahertz, microhenries, and microfarads is useful for rf work. Results will still be in ohms and mhos. Other scaling can be used but should only be done if you're very familiar with impedance and admittance.

## the branch jumping adjacent nodes problem

This situation normally defeats the ladder configuration. The remedy is to use a transformation program on precalculated values until the ladder is formed. **Fig. 5** illustrates the method using a delta-to-tee transformation on a subsection. Circled numbers are nodes above ground. Nodes 1 and 3 will carry over but the original node 2 is reformed.

**Program 4** follows the formulas given in **fig. 5** with rectangular form entry and polar form solutions. Available program size requires more manual steps. Both parts of each delta branch must be entered in the sequence: Input real part, press *ENTER,* input imaginary part, press *R/S* to continue. A manual conversion to rectangular is required at each tee branch solution.

The program accumulates the common denominator of the transformation using the rectangular form with storage in registers R6 and R7. Numerators require multiplication so the polar form is used to reduce program size. The memory register block may seem confusing but the memory contents change as the program progresses.

After $Z_A$ is entered, a direct store is used to begin accumulation. Registers R0, R1 and R2, **R3** both hold the polar form of $Z_A$ while **R6**, and R7 holds the rectangular form. Entering $Z_B$ will make R6, R7 equal to $Z_A + Z_B$ and R0, R1 equal to $Z_A \ x\ Z_B$ via register arithmetic. A direct store places $Z_B$ in **R4**, R5. The numerator of $Z_I$ has been formed in memory. Entry of $Z_C$ will complete accumulation of the denominator and remaining numerators by register arithmetic.

**Steps 31-40** recall the denominator, convert it to polar form and place the solutions in memory. Memory registers R0-R5 are then recalled in order

| STEP | KEY | X | Y | Z | T | REMARKS |
|---|---|---|---|---|---|---|
| 01 | R/S | IMAG A | REAL A | | | ENTER ZA |
| 02 | STO 7 | IMAG A | REAL A | | | BEGIN DENOMIN. |
| 03 | x⇄y | REAL A | IMAG A | | | |
| 04 | STO 6 | REAL A | IMAG A | | | ACCUM.DENOMIN. |
| 05 | g →P | MAG A | PHA A | | | POLAR CONVERT |
| 06 | STO 0 | MAG A | PHA A | | | Z1 NUMERATOR |
| 07 | STO 2 | MAG A | PHA A | | | Z2 NUMERATOR |
| 08 | x⇄y | PHA A | MAG A | | | |
| 09 | STO 1 | PHA A | MAG A | | | Z1 NUMERATOR |
| 10 | STO 3 | PHA A | MAG A | | | Z2 NUMERATOR |
| 11 | R/S | IMAG B | REAL B | | | ENTER ZB |
| 12 | STO + 7 | IMAG B | REAL B | | | ACCUM.DENOMIN. |
| 13 | x⇄y | REAL B | IMAG B | | | |
| 14 | STO + 6 | REAL B | IMAG B | | | ACCUM.DENOMIN. |
| 15 | g →P | MAG B | PHA B | | | POLAR CONVERT |
| 16 | STO 4 | MAG B | PHA B | | | Z3 NUMERATOR |
| 17 | STO x 0 | MAG B | PHA B | | | Z1 NUMERATOR |
| 18 | x⇄y | PHA B | MAG B | | | |
| 19 | STO 5 | PHA B | MAG B | | | Z3 NUMERATOR |
| 20 | STO + 1 | PHA B | MAG B | | | Z1 NUMERATOR |
| 21 | R/S | IMAG C | REAL C | | | ENTER ZC |
| 22 | STO + 7 | IMAG C | REAL C | | | ACCUM.DENOMIN. |
| 23 | x⇄y | REAL C | IMAG C | | | |
| 24 | STO + 6 | REAL C | IMAG C | | | DENOM.COMPLETE |
| 25 | g →P | MAG C | PHA C | | | POLAR CONVERT |
| 26 | STO x 2 | MAG C | PHA C | | | Z2 NUMERATOR |
| 27 | STO x 4 | MAG C | PHA C | | | Z3 NUMERATOR |
| 28 | x⇄y | PHA C | MAG C | | | |
| 29 | STO + 3 | PHA C | MAG C | | | Z2 NUMERATOR |
| 30 | STO + 5 | PHA C | MAG C | | | Z3 NUMERATOR |
| 31 | RCL 7 | IMAG D | | | | |
| 32 | RCL 6 | REAL D | IMAG D | | | DENOMINATOR |
| 33 | g →P | MAG D | PHA D | | | POLAR CONVERT |
| 34 | STO ÷ 0 | MAG D | PHA D | | | Z1 SOLVE |
| 35 | STO ÷ 2 | MAG D | PHA D | | | Z2 IMPEDANCE |
| 36 | STO ÷ 4 | MAG D | PHA D | | | Z3 MAGNITUDES |
| 37 | x⇄y | PHA D | MAG D | | | |
| 38 | STO − 1 | PHA D | MAG D | | | Z1 SOLVE |
| 39 | STO − 3 | PHA D | MAG D | | | Z2 IMPEDANCE |
| 40 | STO − 5 | PHA D | MAG D | | | Z3 ANGLES |
| 41 | RCL 1 | Z1 PHA | | | | |
| 42 | RCL 0 | Z1 MAG | Z1 PHA | | | |
| 43 | R/S | Z1 MAG | Z1 PHA | | | Z1 DISPLAY |
| 44 | RCL 3 | Z2 PHA | | | | |
| 45 | RCL 2 | Z2 MAG | Z2 PHA | | | |
| 46 | R/S | Z2 MAG | Z2 PHA | | | Z2 DISPLAY |
| 47 | RCL 5 | Z3 PHA | | | | |
| 48 | RCL 4 | Z3 MAG | Z3 PHA | | | |
| 49 | GTO 01 | Z3 MAG | Z3 PHA | | | JMP START DISP |

| MEMORY |
|---|
| R0 ZA MAG |
| ZA ZB MAG |
| Z1 NUMER.MAG |
| R1 ZA PHA |
| ZA ZB PHA |
| Z1 NUMER.PHA |
| R2 ZA MAG |
| ZA ZC MAG |
| Z2 NUMER.MAG |
| R3 ZA PHA |
| ZA ZC PHA |
| Z2 NUMER.PHA |
| R4 ZB MAG |
| ZB ZC MAG |
| Z3 NUMER.MAG |
| R5 ZB PHA |
| ZB ZC PHA |
| Z3 NUMER.PHA |
| R6 REAL PART DENOMINATOR |
| R7 IMAG PART DENOMINATOR |

**Program 4. Delta-to-Tee Transformation.**

and the stops display each tee branch magnitude. Tee branch phase angle is in the Y register at each stop. The stop for $Z_3$ is at step **01**. This not only reduces steps but permits an automatic return to start for transforming the next set of delta values.

**Program 4** will work with either impedance or admittance. The only change is from rectangular form to polar form. As an example, enter

$$Z_A = 3 - j4$$
$$Z_B = 4 + j5$$
$$Z_C = 5 + j3$$

The solutions (converted to rectangular) are

$$Z_1 = 2.375 - j0.875$$
$$Z_2 = 1.750 - j1.500$$
$$Z_3 = 1.300 + j2.650$$

Note the difference in signs for the imaginary part.

## transforming the other way

A prime example is the parallel-T circuit. To form the basic ladder a tee-to-delta transformation is required. This places each transformed delta branch in parallel with the other. Transformation formulas are as follows using the branch designations shown in **fig. 5**

$$Z_N = Z_1Z_2 + Z_1Z_3 + Z_2Z_3$$
$$Z_A = Z_N \;/\; Z_3$$
$$Z_B = Z_N \;/\; Z_2$$
$$Z_C = Z_N \;/\; Z_1$$

**Program 5** performs conversion using the above relations.

A difficulty in programming is the common numerator expression. **Program 4** is needed only to form a sum of three quantities for the common denominator. **Program 5** must form the common numerator from the sum of three products. This eats up program steps so **Program 5** is a bit harder to work with.

All entries must be in polar form. This can be done by entering rectangular values and converting manually. Attention must be paid to the magnitude and angle location in the stack. This is especially true at $Z_2$ entry since the Z and T stack registers cannot be disturbed from locations shown.

Registers R0 through R5 store only the individual impedances. The *numerator* product terms *are* formed in the stack, converted to rectangular, then accumulated in registers R6-R7. The $Z_1Z_2$ term is formed when step **12** is reached. $Z_1Z_3$ is complete by step **25**. **Program 5** requires stack register manipulation and more steps as a result of the common numerator expression.

Limitation of HP-25 program size results in solutions stored *inverted* in memory. Input impedance and the direct solution is in admittance but both are

| STEP | KEY | X | Y | Z | T | REMARKS |
|---|---|---|---|---|---|---|
| 01 | R/S | MAG Z1 | PHA Z1 | | | ENTER Z1 |
| 02 | STO 4 | MAG Z1 | PHA Z1 | | | ZC DENOM. MAG |
| 03 | x⇄y | PHA Z1 | MAG Z1 | | | |
| 04 | STO 5 | PHA Z1 | MAG Z1 | | | ZC DENOM. PHA |
| 05 | R/S | MAG Z2 | PHA Z2 | PHA Z1 | MAG Z1 | ENTER Z2 |
| 06 | STO 2 | MAG Z2 | PHA Z2 | PHA Z1 | MAG Z1 | ZB DENOM. MAG |
| 07 | R↓ | PHA Z2 | PHA Z1 | MAG Z1 | MAG Z2 | |
| 08 | STO 3 | PHA Z2 | PHA Z1 | MAG Z1 | MAG Z2 | ZB DENOM. PHA |
| 09 | + | PHA Z12 | MAG Z1 | | | |
| 10 | x⇄y | MAG Z1 | PHA Z12 | | | |
| 11 | RCL 2 | MAG Z2 | MAG Z1 | PHA Z12 | | |
| 12 | x | MAG Z12 | PHA Z12 | | | Z1Z2 COMPLETE |
| 13 | f →R | RE 12 | IM 12 | | | RECT. CONVERT |
| 14 | STO 6 | RE 12 | IM 12 | | | ACCUM. NUMERAT. |
| 15 | R↓ | IM 12 | | | | |
| 16 | STO 7 | IM 12 | | | | ACCUM. NUMERAT. |
| 17 | R/S | MAG Z3 | PHA Z3 | | | ENTER Z3 |
| 18 | STO 0 | MAG Z3 | PHA Z3 | | | ZA DENOM. MAG |
| 19 | RCL 4 | MAG Z1 | MAG Z3 | PHA Z3 | | |
| 20 | x | MAG Z13 | PHA Z3 | | | |
| 21 | x⇄y | PHA Z3 | MAG Z13 | | | |
| 22 | STO 1 | PHA Z3 | MAG Z13 | | | ZA DENOM. PHA |
| 23 | RCL 5 | PHA Z1 | PHA Z3 | MAG Z13 | | |
| 24 | + | PHA Z13 | MAG Z13 | | | Z1Z3 COMPLETE |
| 25 | x⇄y | MAG Z13 | PHA Z13 | | | |
| 26 | f →R | RE 13 | IM 13 | | | RECT. CONVERT |
| 27 | STO + 6 | RE 13 | IM 13 | | | ACCUM. NUMERAT. |
| 28 | R↓ | IM 13 | | | | |
| 29 | STO + 7 | IM 13 | | | | ACCUM. NUMERAT. |
| 30 | RCL 1 | PHA Z3 | | | | |
| 31 | RCL 3 | PHA Z2 | PHA Z3 | | | |
| 32 | + | PHA Z23 | | | | |
| 33 | RCL 0 | MAG Z3 | PHA Z23 | | | |
| 34 | RCL 2 | MAG Z2 | MAG Z3 | PHA Z23 | | |
| 35 | x | MAG Z23 | PHA Z23 | | | Z2Z3 COMPLETE |
| 36 | f →R | RE 23 | IM 23 | | | RECT. CONVERT |
| 37 | STO + 6 | RE 23 | IM 23 | | | ACCUM. NUMERAT. |
| 38 | R↓ | IM 23 | | | | |
| 39 | STO + 7 | IM 23 | | | | ACCUM. NUMERAT. |
| 40 | RCL 7 | IM N | | | | |
| 41 | RCL 6 | RE N | IM N | | | NUMERATOR |
| 42 | g →P | MAG N | PHA N | | | POLAR CONVERT |
| 43 | STO ÷ 0 | MAG N | PHA N | | | YA SOLVE |
| 44 | STO ÷ 2 | MAG N | PHA N | | | YB ADMITTANCE |
| 45 | STO ÷ 4 | MAG N | PHA N | | | YC MAGNITUDES |
| 46 | x⇄y | PHA N | MAG N | | | |
| 47 | STO − 1 | PHA N | MAG N | | | YA SOLVE |
| 48 | STO − 3 | PHA N | MAG N | | | YB ADMITTANCE |
| 49 | STO − 5 | PHA N | MAG N | | | YC ANGLES |

ALL ENTRIES PLACED IN ORDER IN STACK AS SHOWN IN POLAR FORM

MEMORY

| | |
|---|---|
| R0 | Z3 MAG<br>YA MAG |
| R1 | Z3 PHA<br>YA PHA |
| R2 | Z2 MAG<br>YB MAG |
| R3 | Z2 PHA<br>YB PHA |
| R4 | Z1 MAG<br>YC MAG |
| R5 | Z1 PHA<br>YC PHA |
| R6 | REAL PART<br>NUMERATOR |
| R7 | IMAG PART<br>NUMERATOR |

RESTART BY R/S

**Program 5. Tee-to-Delta Transformation.**

in polar form — the input/output is opposite that of **Program 4**. Solutions are manually recalled after the forced stop at step **49**

$Y_A$ magnitude in R0, angle in R1
$Y_B$ magnitude in R2, angle in R3
$Y_C$ magnitude in R4, angle in R5

The admittance form is not a disadvantage. Recalling the parallel-T, the transformations have all delta branches in parallel. Admittance addition is the easiest calculation of parallels.

## other transformations

Lattice networks and similar circuits can be programmed to tee or delta sections and the formulas are available in most textbooks. In most cases a circuit can be divided into relatively simple subsections represented by a single branch; a precalculation program can be used instead of transformation. The circuit to be analyzed should be carefully studied and divided into branches that minimize pre-solution work.

A key indicator of the necessity for transformation is the model's signal path or current flow. The parallel-T has two separate signal paths so transformation is unavoidable. Branch $Z_B$ of **fig. 5** represents a second signal path vs flow from node 1 through node 2 to 3.

Schematics can sometimes be misleading. Often they are drawn to fit available space and parallel components may appear widely separated. Copying the schematic on scratch paper is useful; branches

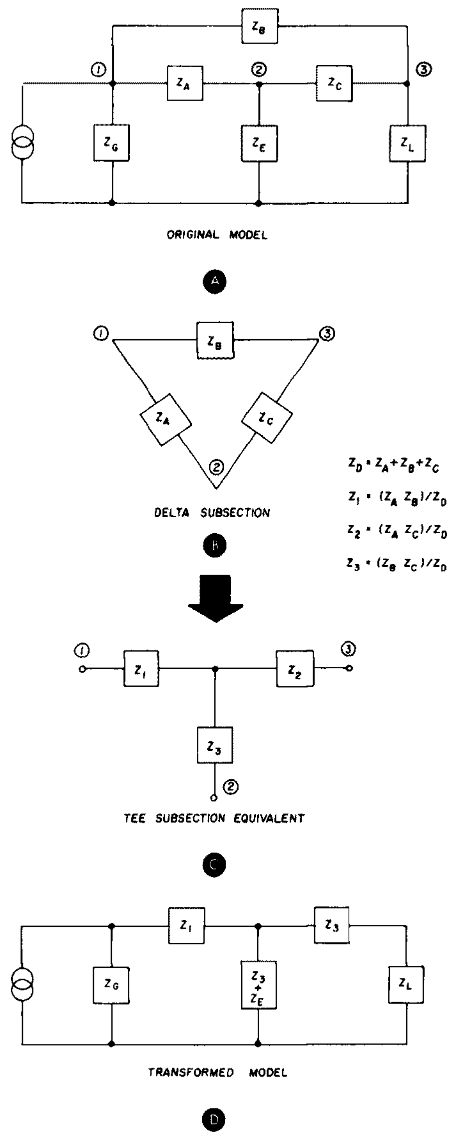



**fig. 5. Model transformation. The delta subsection of the original model is first transformed to an equivalent tee network, then added to the original to form the transformed model.**

can be marked or outlined without harming the original source.

## preparation

Practice with the calculator so you know what happens when you inadvertently key in a wrong instruction. Treat it like a piece of test equipment. In this application that is just what it is.

Use a separate pad of paper for branch value tabulation. When transformation is required, precalculate only the sub-sections and write down the solutions on the tabulation sheet first. The ordinary branches can be calculated later.

Remember to order branch values from back to front. It is probably easier to group values by frequency, making little blocks of values on the tabulation sheet. Calculate at more frequencies than you think you need. Network solutions will tell you which ones to skip over.

Five or six significant digits will give you good accuracy. A minimum is four digits, but only for simple circuits.

## group delay

Any circuit with inductance and/or capacitance has time delay and such delay is frequency sensitive. Group delay is the average time delay over a small increment or *group* of frequencies. It is a function of frequency increment and the differential phase angle within the increment. The formula is:

$$td = (\Delta phase\ angle)/(360 \bullet \Delta Frequency)$$

where: $\Delta phase\ angle$ = positive differential angle in degrees within the frequency increment

$\Delta Frequency$ = differential frequency over a small increment

Scaling can be seconds and Hz, milliseconds and kHz, or microseconds and MHz.

Frequency increments should be constant and small. For a bandpass filter, at least ten frequencies within the passband should give an accurate picture of group delay. Since the frequency increment is constant, the denominator of $td$ is also constant.

Linear group delay in ssb filters will improve audio quality since all sideband components have the same amount of delay. Knowlede of rf to i-f to detector signal delay in receivers is useful when applying noise blankers of the Lamb or Collins type because the blanking signal must arrive before the noise reaches the blanking gate.

## is it worth the trouble?

Only if you do your own circuit designs or want to know what happens inside that collection of components. Programming is not difficult with the HP-25. Hewlett-Packard supplies a good applications manual free and the examples can give a quick insight to the various functions. The same is true for machines made by Texas Instruments.

Computer aided design is used extensively in industry for design analysis. Calculator aided design should be equally useful to the amateur. A few extra hours of *testing on paper* will let you know how the circuit works before cutting metal or etching boards. You can spot errors that would otherwise cost money and time.

**ham radio**


# high-performance 20-meter receiver with digital frequency readout

Design and construction of a dual-conversion receiver featuring varactor tuning in the front end and digital display of the received signal frequency

**Described in this article** is an all-solid-state dual-conversion receiver for 20 meters. It features varactor tuning in the rf, mixer, and high-frequency oscillator and a digital display of the received-signal frequency.

Although the design is for the amateur 20-meter band, it may be easily adapted for a single-band receiver operating in any amateur or shortwave band between 3 — 30 MHz.

Various LC networks and voltage combinations are used with the high-frequency oscillator, which allows limiting bandspread to the CW or ssb portions of a band or expanding bandspread to full amateur-band or shortwave-band segments in the order of 1 MHz.

The receiver portion uses a combination of dual-gate mosfets, fets, bipolar transistors, and ICs. The counter portion uses TTL, low-power TTL, and Schottky TTL logic devices. With suitable selection of HFO and LO frequencies within the capabilities of the Schottky logic upper-frequency limit, the counter may be used up to the 6-meter amateur band with no design changes.

## the receiver

**Fig. 1** is a block diagram of the receiver. It uses the straightforward dual-conversion design approach. Voltage-controlled varactor diodes are used for frequency selection in the HFO and tuning of the rf and mixer input stages. Local and BFO oscillator frequencies are established with crystal control for stability of the i-f. Intermediate frequency amplification and selectivity are obtained with a single integrated circuit in combination with a Collins narrow bandpass mechanical filter (2.1 kHz nominal).

By **M. A. Chapman, K6SDX,** 935 Elmview Drive, Encinitas, California 92024

A simple 5-digit, 7-segment LED display system provides ± 1 kHz accuracy for the receiver frequency display. It uses input signals from all receiver oscillator sections rather than just the HFO.

A nominal position is provided in the BFO selector switch to provide for interpolating the received signal resting carrier frequency. Incorporated into the audio system is an active lowpass filter, which minimizes spurious noise and i-f heterodyne signals, thus optimizing the receiver signal-to-noise ratio.

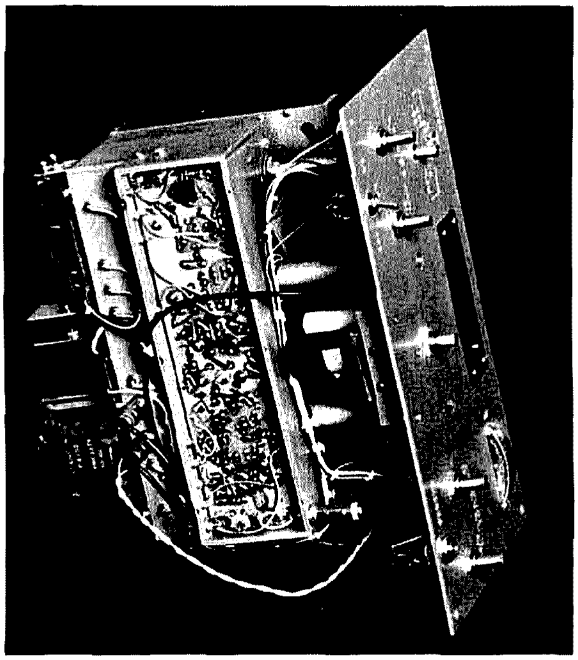
**Frontend. Fig. 2** illustrates the frontend design. Moderately high Q coils are used in the rf and first mixer input. High Q coils in the rf section provide excellent signal-to-noise ratios, good selectivity, and high initial gain. Back-to-back varactor diodes are used for tuning the rf LC circuits by front-panel potentiometer control of the varactor diode bias voltage.

From **fig. 2** we see that the first i-f is 1500 kHz, and the second conversion i-f is 455 kHz. As shown in **fig. 2**, the gates of the second mixer stage incorporate bias adjustment to optimize both device gain and signal-to-noise ratios. Gate biasing for rf and mixer may not be immediately apparent from **fig. 2. Fig. 3** shows how the mosfets in the rf and mixer biased and the method used to compensate for tolerances in the LC circuits to permit gang tuning of both stages.

The upper section of **fig. 3** is a simplified ac model illustrating the tuned-circuit relationships. It's important to note that, from an rf standpoint, the cold end of the LC circuits is at ground potential as are the source and control gates of both devices. The lower section of **fig. 3** includes the varactor diode and source-swamping-resistor relationship to the device. Here again, we see that the cold ends of the LC circuits are at ac ground and that the source swamping resistor is shunted to ground.

Since the rf amplifier and mixer are depletion mode devices, you can adjust the input voltage to the gates of Q1 and Q2 so that the gate-to-source bias is approximately − 0.5 volt. Tuning the varactor diode is accomplished by applying a positive voltage greater than that required for bias of the gate source

Author Chapman has provided much detail in his description of the design and construction of this receiver. The article is in two parts. Part one describes the basic rf and i-f circuits, and part two deals with the digital readout system. The digital readout system may be readily adapted to almost any receiver design. If you're interested in duplicating the receiver without the digital readout, you'll find the information you'll need in the following material. Part two, to be published in a subsequent issue of *ham radio*, wraps up the entire project and provides much useful information on calibration and alignment. Editor.

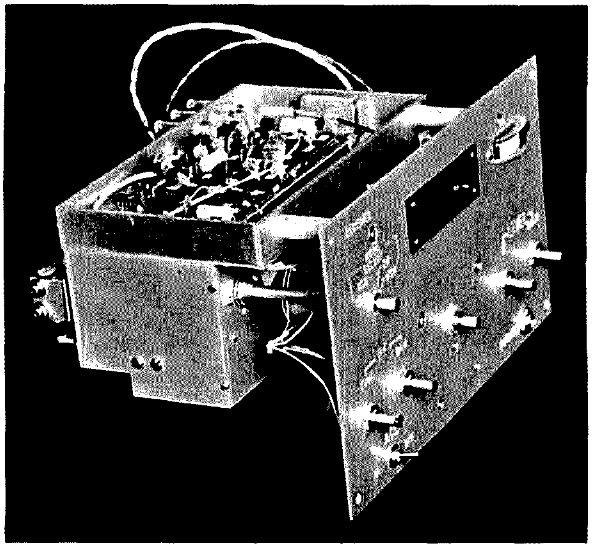


**Bottom view of the receiver. A portion of the power supply board can be seen on the left.**

terminations. The varactor must be reverse-biased for it to operate in a nonconductive state and act as a variable capacitance across the inductors. For initial setting of the rf and mixer stages, bias voltage no. 1 should be adjusted for 0.5 volt *differential* bias to the source voltage point no. 3, with the control gate at maximum voltage (approximately 8 volts dc).

The initial bias and source-point voltages will be approximately +0.75 volt. Examining the voltage divider network in the varactor diode tuning circuit, we see a minimum positive voltage of approximately 2.25 Vdc, providing a 1.50-volt reverse bias in the worst case and an 8 Vdc bias in the high case. You can optimize the voltage-divider circuit after assembly and test to increase or decrease the bandspread capabilities. Tuning the 20-meter band will require about 1.5 volts of differential. If you wish, optimizing can be implemented, so that the differential occurs on the major portion of the potentiometer adjustment range.

To minimize alignment difficulties, Q1 and Q2 should be matched for nominal $I_{DSS}$ values, so that bias voltages 1 and 2 will require a minimum amount of adjustment to compensate for varactor diode and shunt-capacitance tolerance. Much variation exists between varactor nominal capacitance and capacitance change, with respect to voltage change when tuning the rf circuits. Compounding the problem is the nonlinearity that exists with the diode elements themselves. Since the Q of the rf input and mixer input circuits is quite high (~75), any differences be-



**Side view of the receiver showing the rugged mechanical construction.**

tween the diode-versus-voltage capacitance will result in circuit detuning.

Referring to **fig. 2**, the mixer gain control gate (G2) is at a fixed-voltage point and the rf stage uses a variable resistor to control the rf gain. One disadvantage of this scheme is that, as we vary the voltage

limited segments of a band (for example, the CW section), then very little rf **TUNE** adjustment is required. However, if a number of changes are made to the mixer gain, a corresponding adjustment to the **TUNE** control is necessary.

**I-f amplifier.** I-f amplifier gain and bandpass filtering is accomplished with an LM373H IC and a narrow bandpass Collins mechanical filter (**fig. 4**). The LM373H (U1), in addition to providing high i-f gain, also acts as the detector for both CW and ssb. Incorporated in its design is the ability to use agc or external manual i-f gain. An agc-level sensitivity control is included.

The voltage value indicated for initial adjustment of U1 represents an agc threshold setting of approximately 300 microvolts. To provide isolation of the detected audio output to the S meter and audio filter stages, a simple fet amplifier stage is incorporated (Q4). Almost any depletion-type device may be used in this stage by adjusting the source and drain resistors to obtain a voltage gain of approximately 5, as indicated. The minimum source bypass capacitance should be in the order of 10 $\mu$F.

The Q4 output coupling capacitor should provide a low reactive impedance to the S meter and audio filter input to minimize signal loss at the low (150 Hz) signal levels. However, scaling of the S-meter input

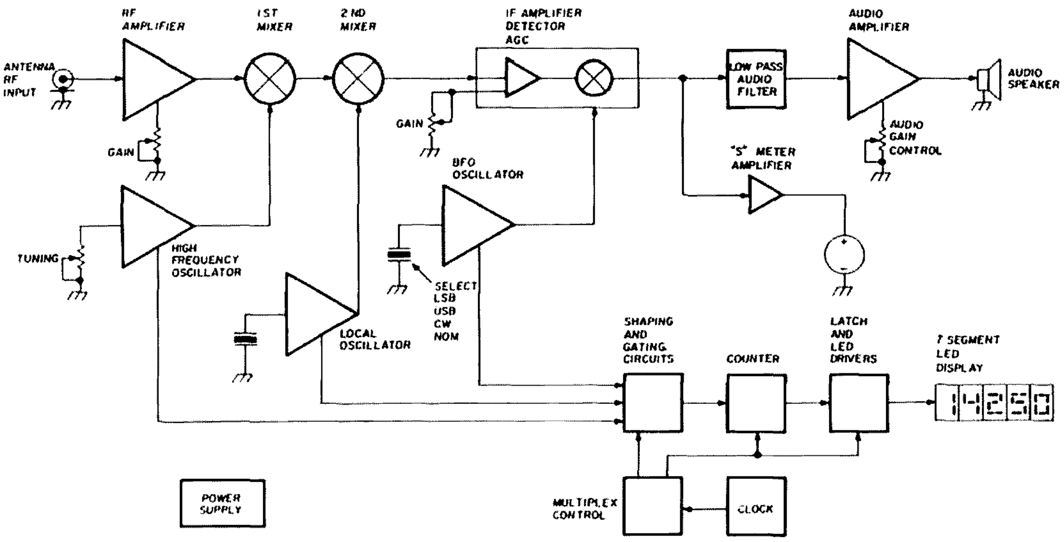



**fig. 1. Block diagram of the 20-meter high-performance receiver.**

of Q1 gate 2, causing a change in the varactor diode bias and its capacitance, a slight detuning occurs in the LC circuits.

When the gain level is adjusted to any appreciable extent, then the **TUNE** control must be repeaked to maintain overall-receiver gain. When operating over

is possible, since at this point we are only interested in the audio voltage level — very little current is drawn by this or the succeeding stage.

The S-meter amplifier is fully described in reference 1. The builder may elect to use either of the S-meter circuits described in the referenced article

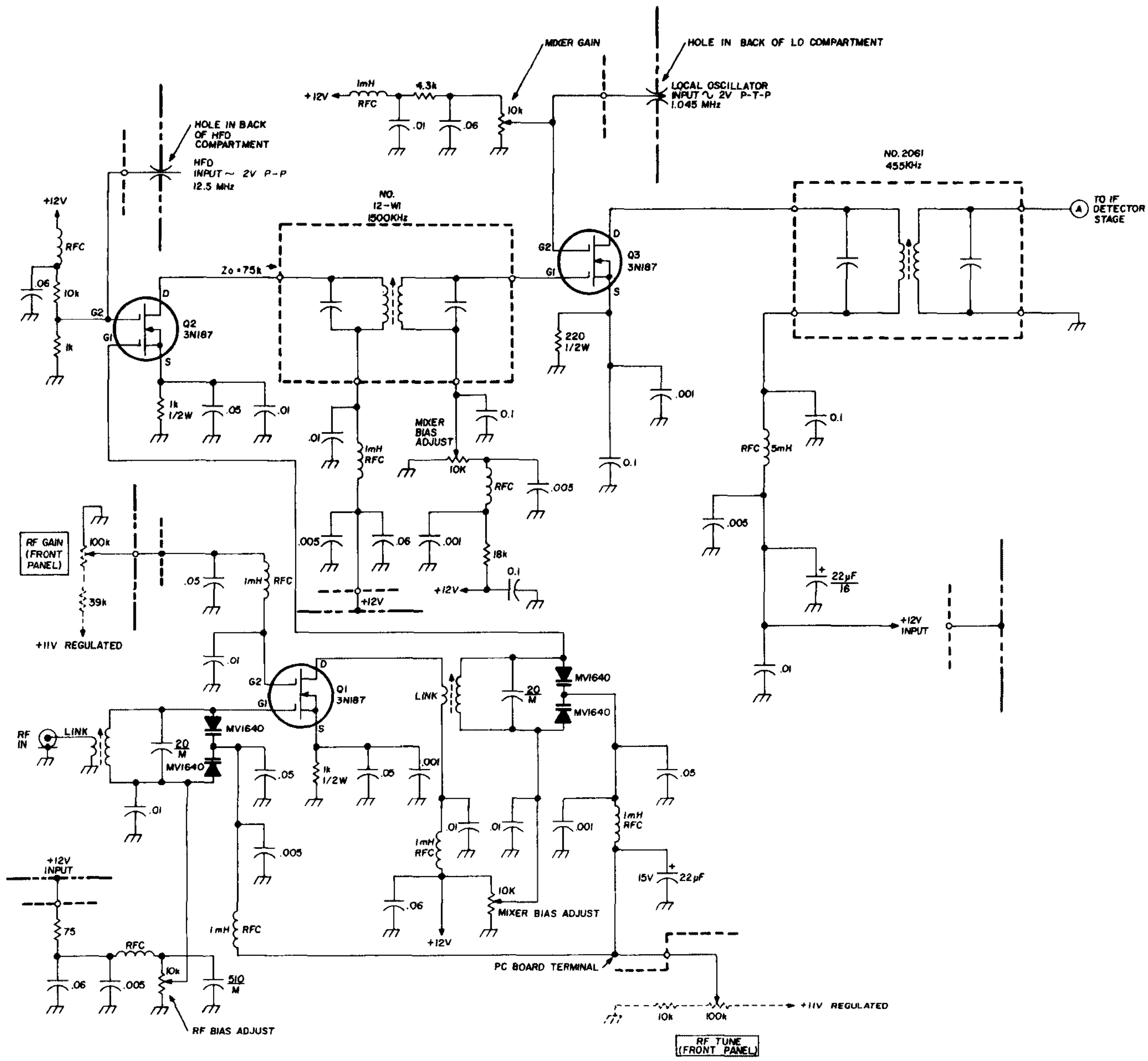


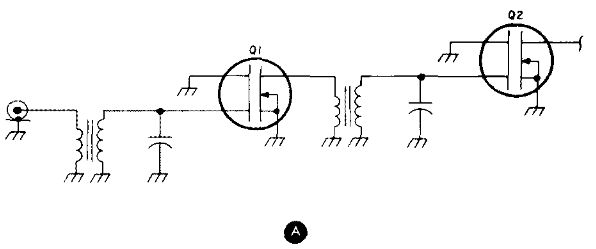

fig. 2. Rf and mixer schematic.

depending upon panel layout preference, and availability of meters.

**High-frequency oscillator.** An extremely high stability HFO is illustrated in **fig. 5**. This is an adaptation of the design discussed in reference 2 using a Goral buffering scheme. I've used this oscillator technique from 1 to 50 MHz with various LC combinations, changing only the source and emitter load inductors. An alternative device providing equal or better output voltages is the Motorola MPS-A09 bipolar transistor. Since the mixer-stage gate and counter input circuitry are high impedances, very little signal loss occurs, and you should expect to see the output levels indicated on the diagram.

Substitution of alternative inductive loads can affect the output voltage levels. Use care to obtain the peak-to-peak values indicated in all oscillator stages.

Tuning of the varactor (VVC, MV1654) is accomplished using a positive voltage of approximately 6 to 7.5 Vdc. Since the first i-f is 1500 kHz, and down conversion is used, the HFO output, then, is between 12.5 MHz - 12.85 MHz to cover the 20-meter band (14 - 14.35 MHz). Considering that 1.5 volts of differential bias on the varactor is necessary, the fre-



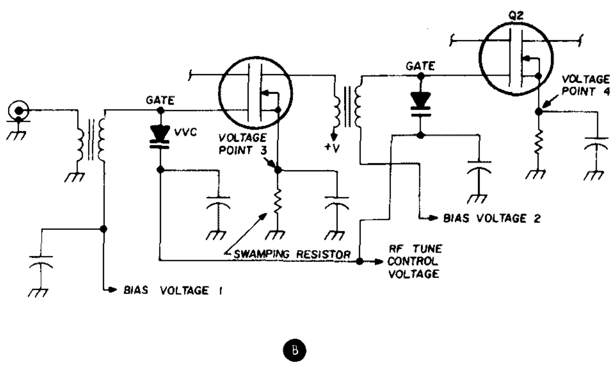




fig. 3. Rf amplifier tuning and bias schematics. Sketch A is a simplified ac model illustrating the tuned-circuit relationships. Sketch B includes the varactor diode and source-swamping-resistor relationships.

quency stability is more sensitive to power-supply ripple variations than the oscillator described in reference 3; therefore, a high-performance voltage reference supply is necessary to maintain a stable voltage on the varactor to minimize frequency drift. Presuming that the power-supply scheme illustrated later is used, and that good packaging and component selection is followed, you can expect a frequency drift of less than ±100 Hz/hour after thermal equilibrium.

***Local and beat-frequency oscillators***. The local and beat oscillators are described in **figs. 6** and **7** respectively. Simple fet oscillators are used. Many fet substitutes will perform equally as well. The output voltages are as indicated on the schematic. Good-quality parallel-resonant crystals should be used, with only the nominal 455-kHz crystal requiring careful selection of the shunting capacitor, since this will determine your transmit-frequency reference in the counter display.

Any errors introduced into the local-oscillator center frequency can be compensated by shifting the HFO frequency and tweaking the 1.5-MHz i-f transformer (**fig. 2**).

Caution note: the LM373H (U1) of the i-f detector stage is sensitive to excessive BFO injection voltages. The nominal input voltage to U1 for BFO detection is 60 millivolts rms. Slightly higher levels will *increase* the detected audio levels. A too-high injection level will result in audio motorboating and distortion; too low an injection level will result in poor or weak audio detection. To minimize generation of spurious and harmonic frequencies from feeding into other parts of the receiver, generous use is made of RC decoupling and bypass capacitors. Feel free to substitute capacitor values in these circuits with

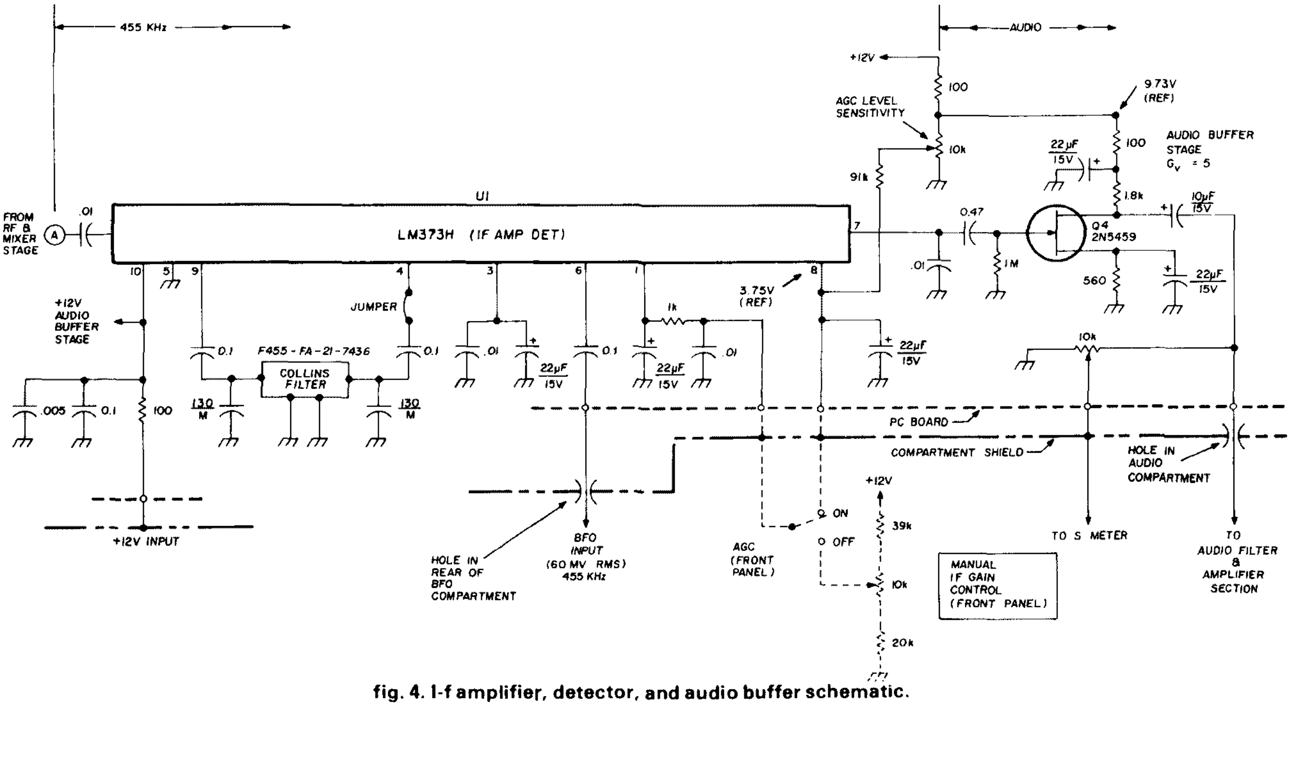



fig. 4. I-f amplifier, detector, and audio buffer schematic.

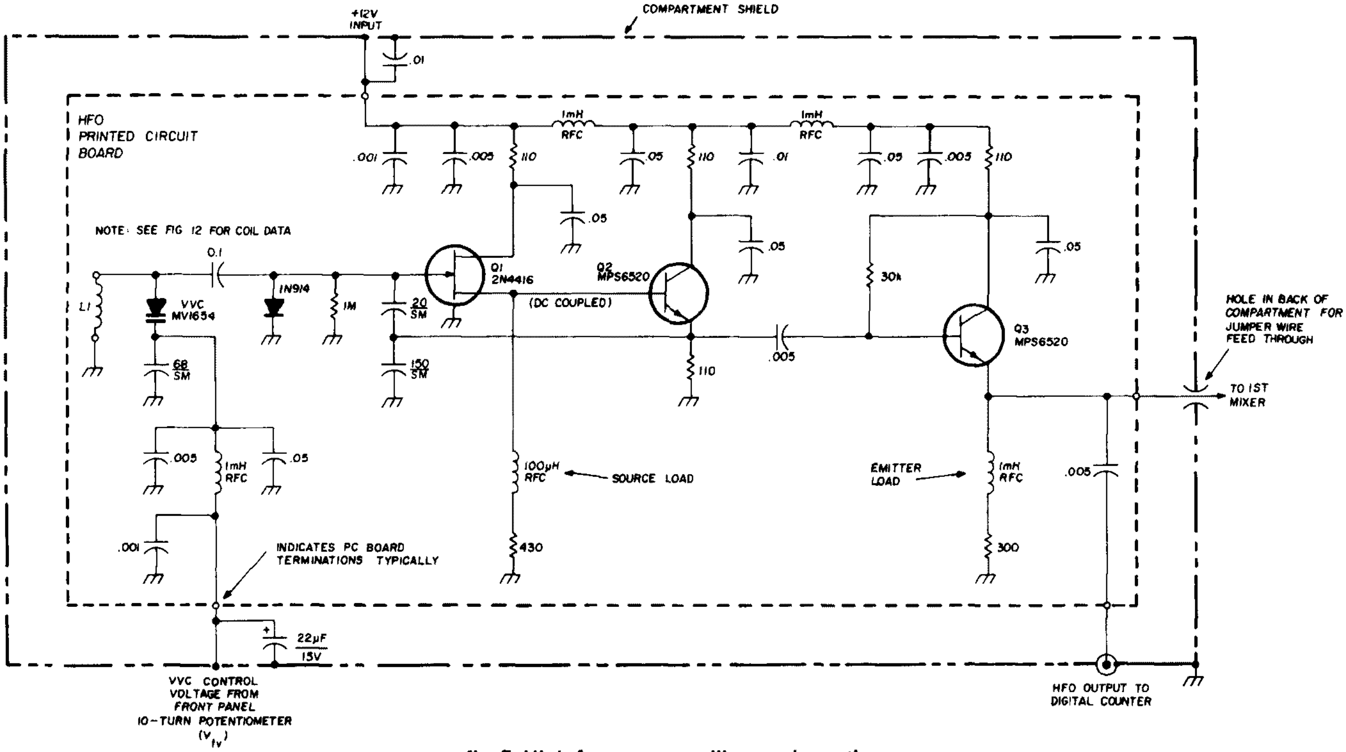



**fig. 5. High-frequency oscillator schematic.**

whatever is available, so long as the dc input is kept reasonably clean of the oscillator-frequency components.

The LSB and USB crystal frequencies may be determined by either measuring the Collins 455-kHz mechanical filter for the 20-dB passband points or by ordering a filter that has been measured by the manufacturer for the L20 and H20 dB points. An additional description is given in reference 4 and in the alignment and calibration sections of the article.

**Audio filter and amplifier.** The detected audio from Q4 is passed through an active lowpass filter-operational amplifier arrangement and further amplified in a one-stage 2-watt audio amplifier U3 (fig. 8). The U1 and U2 lowpass filter design is discussed in considerable detail in references 4 and 5. Normally, U1 and U2 operate from a plus and minus voltage source; however, a simple voltage-divider circuit on pin 3 of U2 establishes an artificial ground for both U1 and U2. The lowpass rolloff starts at 2500

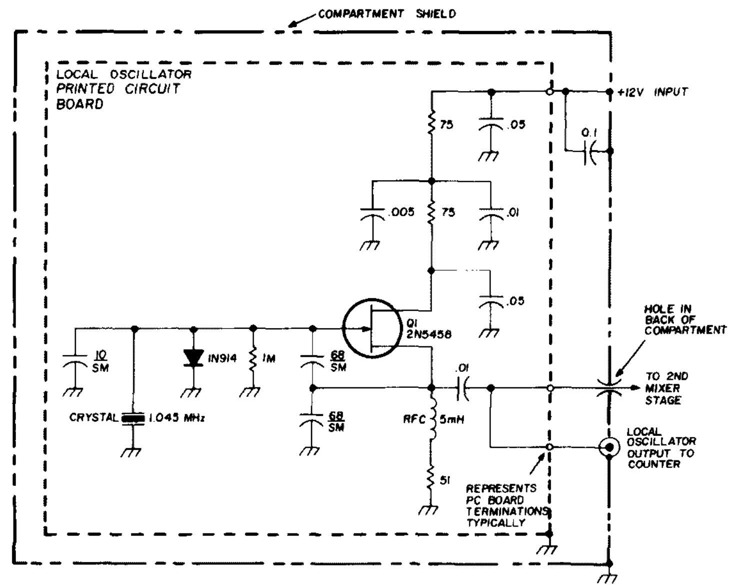



**fig. 6. Local-oscillator schematic.**

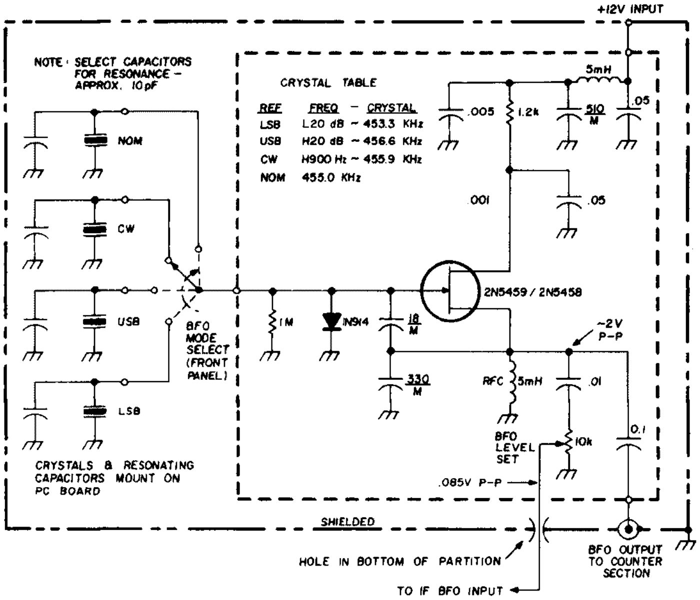



**fig. 7. BFO schematic.**

Hz, and a 15-20 dB attenuation of all higher audio frequencies is achieved. The normal i-f heterodyne hiss is greatly attenuated, and the receiver overall signal-to-noise ratio is enhanced. An additional high-frequency audio rolloff is achieved by the selection of the 680 pF mica feedback compensating capacitor of U3.

Audio and rf have many common grounding and feedback problems. To minimize circulating ground loops and achieve low-noise performance, a single-point ground system should be used with the audio circuit. As an aid, those items of special significance are indicated as notes in **fig. 8.**

Notes 1 and 2 are self-explanatory; note 3 deserves some additional comment. A 1-meg level-set pot is shown between U2 pins 2 and 7. As the name implies, it establishes U1 and U2 output gain in combination. Ideally the gain is unity; however, since we're operating these devices ±6 volts dc the overall gain will be in the order of 0.8-0.9. To obtain the best

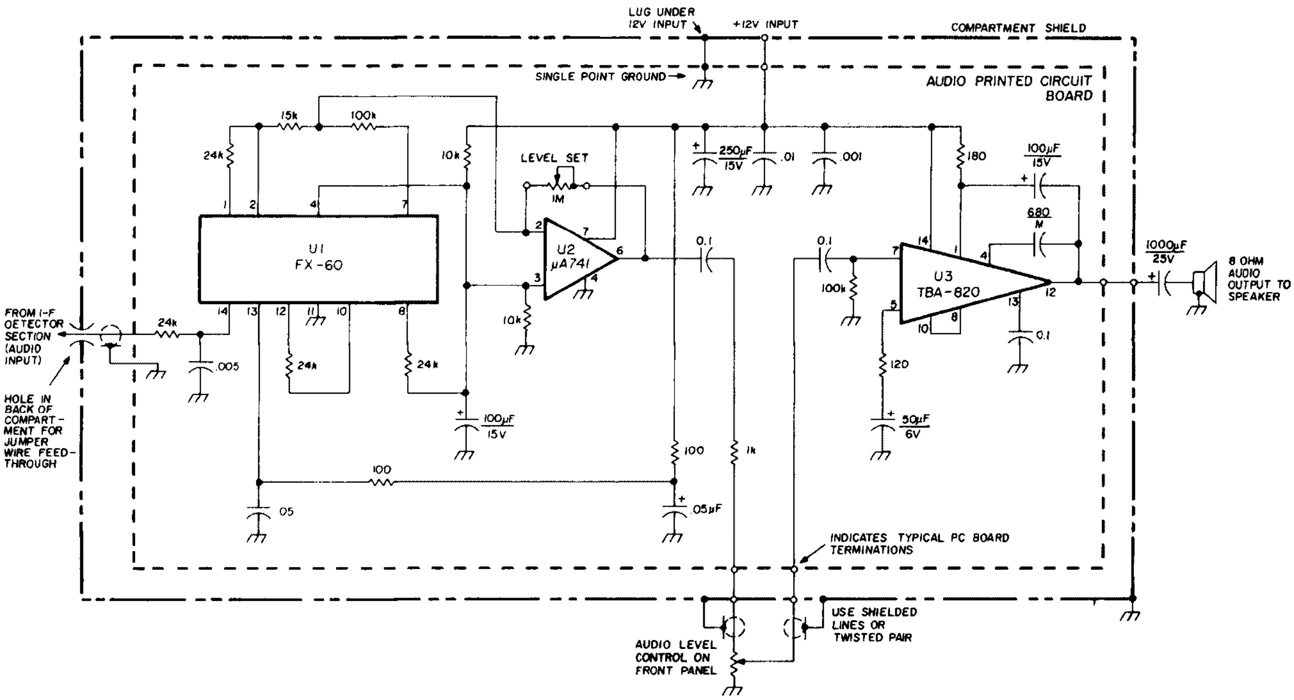



**fig. 8. Audio filter and amplifier schematic.**

setting for the level set, a low-level 50-100 millivolt audio signal should be coupled to U1 input through a 0.01 $\mu$F capacitor while monitoring U2 and U3 output with a high-impedance scope. The level-set pot should be adjusted where the gain is about 90 per cent of maximum over the frequency range 150 Hz-2500 Hz. Then the pot may be removed and a fixed resistor installed with a value closest to that measured on the pot. The value of this resistor will be approximately 100k.

As in the other low-level audio-signal processing circuits, this is not the place to use an inexpensive, exposed wiper-type volume control. The added cost of a sealed, 2-watt type component with the case securely grounded to the chassis and panel will minimize wiper contact noise and ground problems. Use of 1/2-watt resistors in the low-level audio stages has been included in the PC design layout for additional noise control.

**Power supply.** The power supply and high-frequency oscillator varactor control system are illustrated in **fig. 9**. Three-terminal 7800-series voltage regulators are used for the +12 and +5 Vdc requirements. To minimize power dissipation and provide for a constantly changing current demand from the LED displays, a preregulator/derippler circuit is used ahead of the +5 Vdc regulator. A small amount of series resistance isolates the display voltage from the +5 volt logic voltage.

To establish a stiff reference voltage for the HFO varactor diode ($V_{FV}$), a precision voltage regulator is used (U2). Ripple rejection is in the order of 80 dB, and voltage stability is in the order of a few millivolts per degree F. Two small 10-turn trimpots allow fine tuning of the high and low end of the 10-turn $V_{FV}$ voltage pot. A similar voltage control system is described in reference 3. Additional capacitive filtering will aid in the reduction of $V_{FV}$ ripple content and may be added at your discretion.

Filtering across the $V_{FV}$ 10-turn pot should not be added, because such filtering will increase the HFO control-voltage time constant. Any added filtering should be used ahead of the trimpot or preferably U2.

## construction

Modular PC board construction is used, with each major function individually packaged on a separate board and isolated in an rf-tight housing. The photos clearly illustrate the relative board positions and techniques for interfacing the rf and audio circuits, as well as the method for isolating and developing the +12 source voltage between the receiver stages. Each functional section has its own +12-volt input

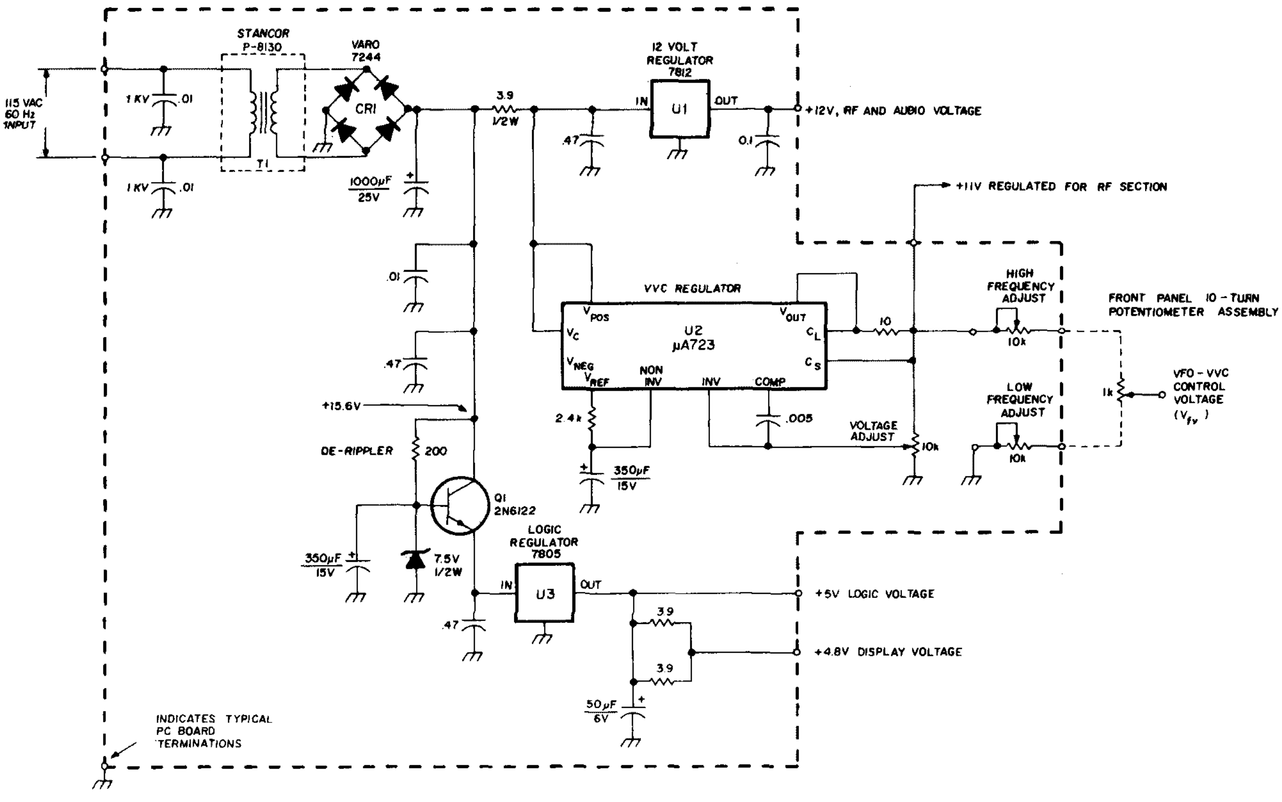



fig. 9. Power-supply schematic.

terminal, which is filtered at the board termination point.

**Front end and i-f amplifier/detector.** The rf, mixer, i-f and detector functions are located on one inline PC board (**fig. 10**). Heavy, well grounded shields are used between the rf, mixer, and i-f/detector sections, which also provide convenient mounts for the large coils of the rf circuits. Two standard, readily available i-f transformers are used in the drain circuits of Q2 and Q3, which are shielded with their own housings. Many 455-kHz i-f transformers having the same or similar geometry as the 2051 unit used here are available. If necessary, the board could be cut open and the alternative unit mounted in piggyback style.

If you'd like to operate the receiver at frequencies above 15 MHz, the 12-W1 transformer may be opened and the internal padding capacitors changed to allow for a higher first i-f more consistent with a good 10:1 rf mixer frequency ratio. If you elect to use an unprotected dual-gate mosfet device in the rf stage, provisions exist for the installation of back-to-back 1N914 gate-protection diodes; however, at higher frequencies this additional shunt capacitance could materially affect the signal level.

The Collins mechanical filter mounts directly to the board. However, several foreign-made filters of comparable performance are available, which will mount in this area of the board. Short jumper wires can be installed to interface with the input-output pins of the LM373H. Provisions are included on the board for an impedance-matching resistor to be installed in place of the jumper wire input to U1 and the i-f detector from the filter.

Inexpensive 0.1-inch (2.6mm) center trim pots were used; however, after measuring the desired resistance values required for the S-meter, agc, and bias circuitry, fixed carbon composition resistors mounted vertically to the board could be used.

**High-frequency oscillator.** The high-frequency oscillator PC component layout is shown in **fig. 11**; **fig. 12** shows the winding and coil data. As discussed in reference 3, a variety of LC, varactor-diode combinations are possible for this construction. Provisions exist for mounting two varacaps back-to-back on the board, and space is available for the addition of an extra shunt capacitor. The HFO output signal to the mixer should have the shield terminated on the HFO board and should be unterminated at the mixer input. Note that the board should be *well-grounded* to the chassis. Since the $V_{FV}$ voltage is at a dc level, there should be no need to use shielded wiring unless an alternative packaging technique required different routing of this wire.

**Local oscillator.** The local oscillator is mounted on an independent PC board, as shown in **fig. 13.** Wiring and installation is straightforward. Many crystal sizes can be accommodated on this board. Provisions are included for an optional ceramic or air type trimmer capacitor. However, an error of 2 or 3 kHz is not harmful since this error can be compensated for in the final HFO trimming and adjustment. Crystals with low ESR (sluggish) will not operate well in this circuit and may have a slight tendency to drift, so purchase good-quality crystals or use an alternative circuit similar to the oscillator circuit described in reference 4.

**Beat-frequency oscillator.** The beat-frequency oscillator is also mounted on an independent PC

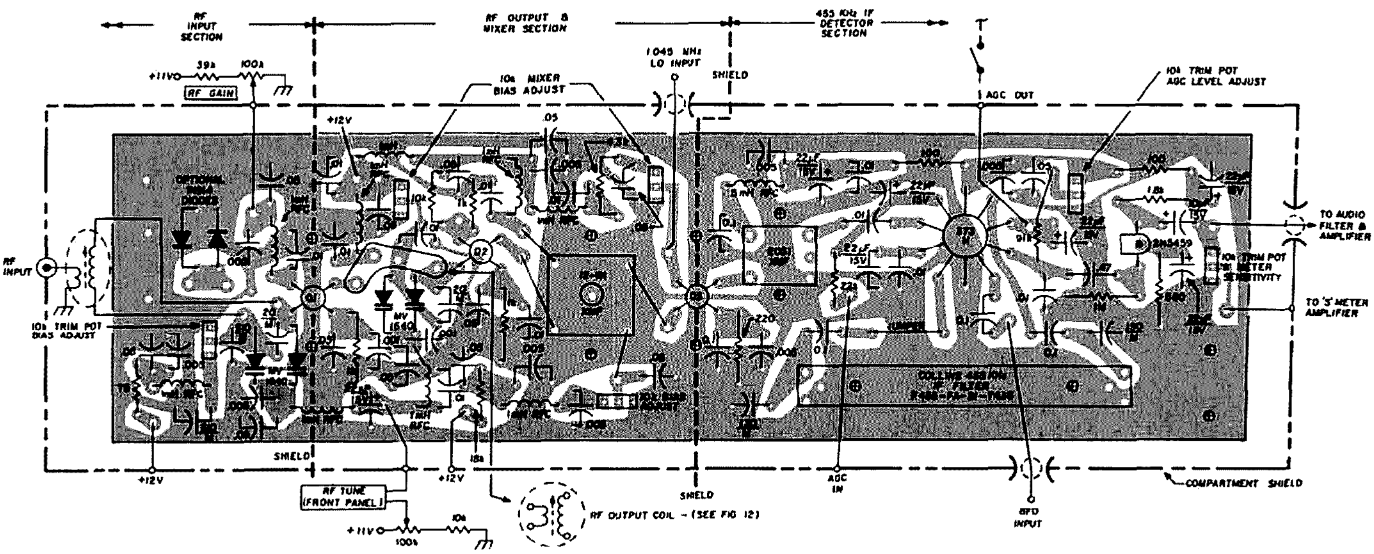



**fig. 10. RF amplifier, mixer, and i-f/detector PC-board assembly. A complete set of circuit board layouts is available by sending a self-addressed, stamped envelope to *ham radio*, Greenville, New Hampshire 03048.**

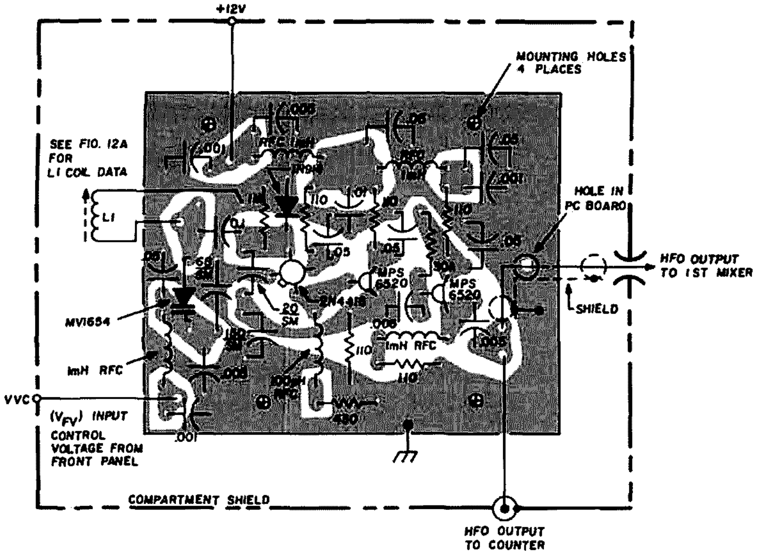



**fig. 11. HFO PC board layout. Printed circuit boards for the rf portion of the receiver are available from the author for $11.25, ppd.**

board (fig. 14). The board assembly mounts directly to the switch mounting hardware behind the switch deck. The switch is a Centralab PA-3 with a PA-2 nonshorting 2-6-position deck. Wiring from the PC board to the switch terminals is illustrated in the figure.

As in the local-oscillator board, the crystals are soldered directly to the board; either standard or wire-type crystals may be used. I recommend that all crystals be of the same type to minimize any temperature variations and to provide equal output levels to the counter and mixer input circuits.

As previously mentioned, the USB and LSB crystals depend on the filter 20-dB passband points. With foreign mechanical filters, my experience has been that the filter center frequency may not be at the nominal 455-kHz resting point, but may vary as much as ±5 kHz from this point. If an alternative i-f filter is used, the center-frequency response and that at 20-dB points should be measured as accurately as possible for optimum receiver performance.

The CW crystal is normally ±800-1000 Hz from the center i-f filter frequency. It should be selected according to your preference. Use an audio oscillator to determine the preferred beat oscillator note, when ordering a corresponding BFO CW crystal.

A unique calibration scheme is described in part 2 of this article, which uses the CW and BFO frequency. During tests of the BFO, an accurate measurement of the CW BFO oscillator frequency should be

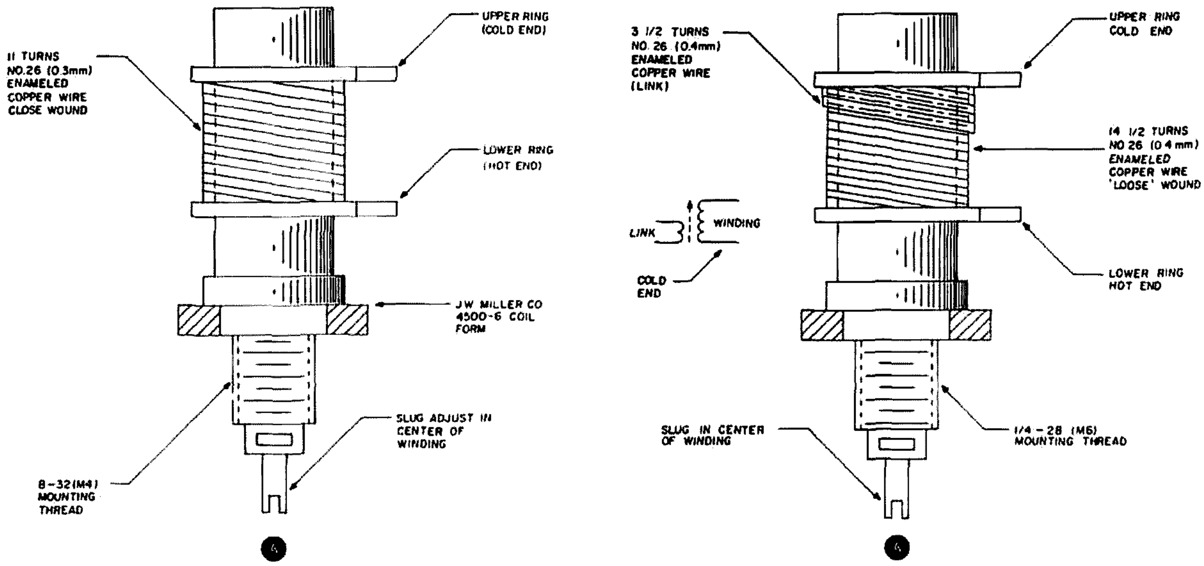



**fig. 12. Construction details for the hf oscillator coil, A, and rf and mixer coils, B.**

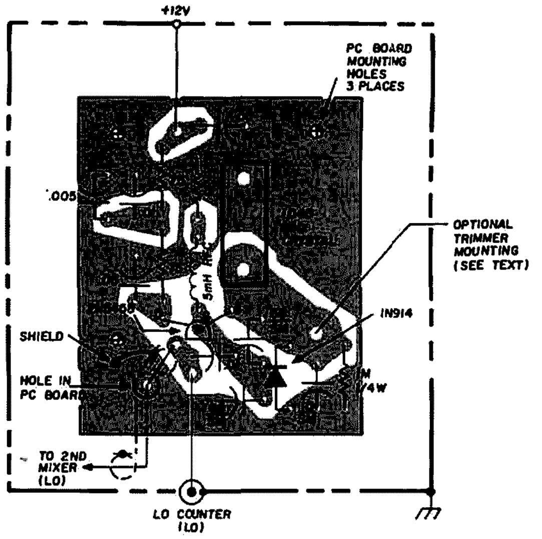



**fig. 13. Local oscillator PC-board layout.**

made if the crystal frequency is other than ±1000 Hz from the filter center frequency. This will allow you to adjust the clock frequency accurately in the counter for the correct display. Both the local oscillator and BFO oscillator PC assemblies should be hard mounted to the chassis, using short aluminum standoffs for good grounding.

**Audio filter and amplifier.** The audio filter and amplifier are included on one PC board assembly, **fig. 15**. Terminals from U1 and U2 must be removed from the integrated circuits before installation. The FX-60 device may be mounted to a 14-pin DIP socket or soldered directly to the board. Grounding within the audio section is important, so a single-point grounding scheme illustrated in the component layout of **fig. 15** is recommended. The PC board should be isolated from the chassis using nylon hardware. The audio input shielded lead should be grounded at the PC board input and left floating at the detector output. Because of the large size of the audio output capacitor, I recommend that this component be located externally, near the speaker or speaker output terminal.

**Power supply.** The power supply component layout is shown in **fig. 16**. Heatsinking of Q1, U1 and U3 is mandatory because of the power dissipation of the regulator devices. The power supply photograph illustrates the technique I used, wherein a small piece of aluminum sheet was bent up, with holes drilled in the side and bottom. Since Q1 collector is not operating at ground potential, its case must be insulated from the heatsink. However, the common terminals of U1 and U3 are at ground potential, and their cases may be tied directly to the heatsink. A small amount of silicone grease under the mounting surfaces of these components will aid in thermal conduction. The peak power dissipation values are indicated on the schematic (**fig. 9**), as a reference for the builder to size the heatsink.

A small aluminum angle bracket is mounted on the end of the PC board assembly to support the HFO trimming resistors; however, these parts can be located at any convenient point on the chassis or next to the 10-turn HFO control on the front panel. The precision voltage regulator, U2, may be mounted to a standard 14-pin DIP socket or soldered directly to the board. Do not forget to add the short jumper wire shown on the back side of the PC board layout of **fig. 16**.

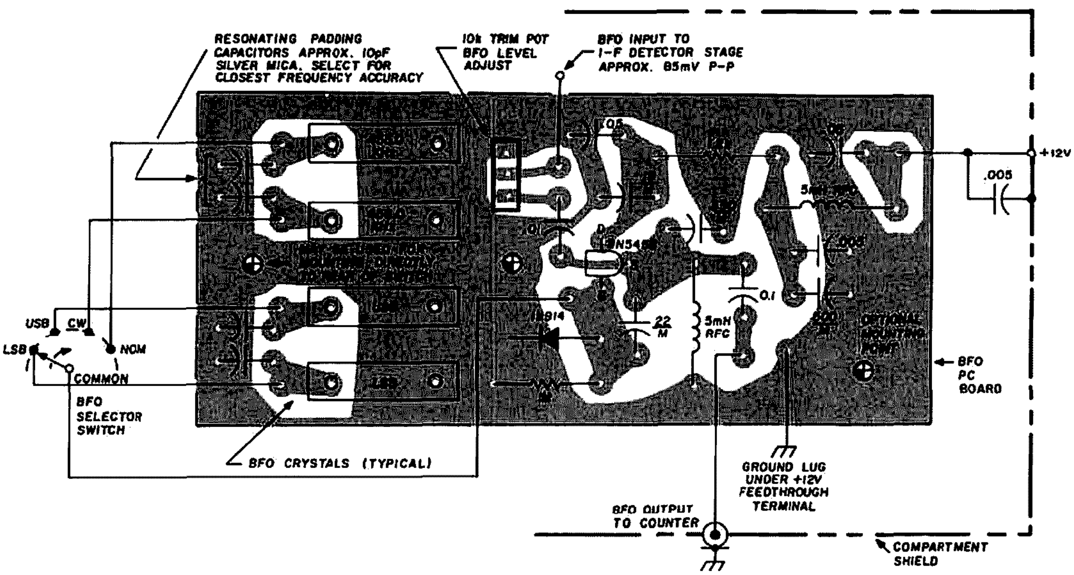



**fig. 14. Beat-frequency oscillator PC-board component layout.**

A variety of bridge rectifier units are available that match the Varo part number 7244 unit; it is important to have adequate surge current capability (approximately 5 amps) in using subsitute units. Four independent rectifiers may be mounted vertical to the PC board face and jumper wired into a bridge arrangement if desired.

## alignment and testing

The BFO i-f bandpass frequency relationship is illustrated in **fig. 17**. This receiver uses a high-grade mechanical filter having a 2.1 kHz bandpass at 60 dB down. Because of this sharp filter action, the location of the ssb beat oscillator frequencies is quite critical. As illustrated, the upper and lower BFO ssb frequencies are precisely located at the 20 dB points. Those ordering the Collins filter part specified in the schematic will receive the unit with the 20-dB points measured at the factory, from which the approximate crystal can be ordered. Since a variety of Collins 455-kHz mechanical filters are available from surplus markets a stable single generator, scope, and frequency counter should be used to measure accurately the beat frequency points and resting center frequency for determining the appropriate BFO crystals.

A complete description of the S-meter circuit, construction and implementation, is described in reference 1. The photographs illustrate the method of S-meter, and S-meter amplifier mounting. In this application I am using an 0-5 mA S-meter movement.

**Preliminary procedures.** The following steps for alignment of the basic receiver section will assist you in obtaining maximum performance. Before alignment, each oscillator and audio module should be tested and verified for proper operation, as discussed in the text. The modules should be installed in their final form on the chassis with appropriate interconnect wiring. If possible, two separate external 0-15-volt power supplies should be used.

**Alignment.** The following steps have been verified in practice.

**1.** Temporarily install one end of a 0.01-$\mu$F capacitor to Q3 gate on the foil side of the rf and mixer PC board. With an external signal generator, inject an unmodulated 455 kHz 1 millivolt signal to the other side of the 0.01-$\mu$F capacitor. Put the agc switch in the OFF position and adjust the i-f gain control for maximum. Verify that the BFO injection signal is in the order of 60 millivolts rms in the CW position.

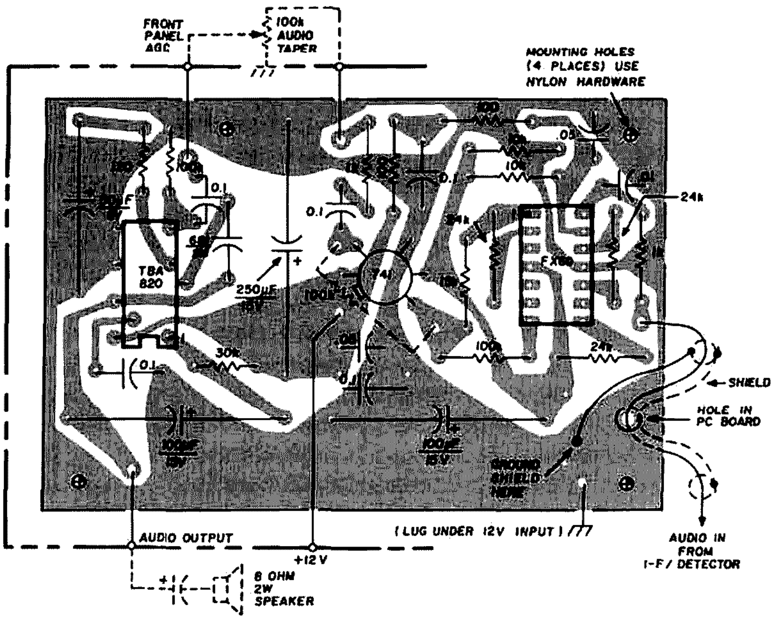



**fig. 15. Audio filter and amplifier PC-board layout.**

Adjust the volume control for maximum and install a scope monitoring probe at the audio output, *without an external speaker installed.* Use caution to ensure that the audio *is not* grounded, since this may destroy the audio amplifier chip!

With power applied ( + 12 V), adjust the 455 kHz i-f transformer in Q3 drain circuit for maximum while monitoring the audio output signal. With the indicated 1 millivolt 455-kHz signal applied to G2 of Q3, a 2.5-volt peak-to-peak audio output signal should be observed.

With the agc set at the level shown ( + 3.75 V), switch the agc control from OFF to ON. Very little change in the detected audio should be apparent.

Remove the temporary 0.01-μF capacitor from G2 of Q3.

2. Temporarily install a 0.01-μF capacitor to Q2 gate 1 on the foil side of the board, as in step 1. With external controls set in the positions as previously indicated, apply a 1.5 MHz CW signal of 1 millivolt to the 0.01 μF capacitor.

Adjust the mixer bias controls for Q2 and Q3 and the 1.5 MHz interstage transformer in the drain circuit of Q2, with power applied. Monitor the audio output as in step 1. With the signal level indicated, a minimum detected audio output of 6 volts peak-to-peak should be observed. Remove the 0.01-μF capacitor from Q2 gate 2.

3. Temporarily install a 0.01-μF capacitor to the drain terminal of Q1, foil side. With the front panel controls set as previously described, apply +12 V power and approximately +6.5 volts on the VVC $V_{FV}$ voltage point of the HFO module. Apply a 1 millivolt signal of 14.125 MHz to the temporarily installed capacitor and adjust the rf gain and if tune controls for maximum detected output. Adjust the rf bias pot on G1 of Q1 so that reverse biasing of the tuning diodes is ensured. With a CW signal applied as described, and all bias adjustments optimized, a 4.5-volt minimum peak-to-peak signal should be obtained at the audio monitoring point.

Remove the 0.01-μF capacitor from the Q1 drain.

Note that some minor adjustment of the +6.5-volt $V_{FV}$ voltage will be necessary to obtain conversion of the injected signal but should be somewhat less than 8.0 Vdc if the HFO oscillator slug is properly positioned.

4. Install a 0.01-μF capacitor on the antenna input link of the rf input tuned circuit. Position all controls as previously indicated, apply power and VVC $V_{FV}$ control voltage. Inject a 30-μV signal at 14.125 MHz, unmodulated, to the input capacitor. Adjust the $V_{FV}$ voltage for obvious beat-detected audio output. Peak the rf tuning and bias adjustments along with the first and second i-f transformers. Repeaking of these controls should be done several times to optimize the detected audio signal.

With a 30-μV signal applied as indicated, a clear detected audio output of approximately 7 volts peak-to-peak should be observed. Reduce the input signal to 0.5 μV. A 1.0-p-p signal should be observed at the audio output.

Temporarily ground the input to the receiver and observe the audio output, which should be in the order of 0.25 volt peak-to-peak while varying the $V_{FV}$ voltage from +4 to +12 volts. There should be no observable birdies or spurious audio output spikes over this frequency translation of the HFO. Using these values, the noise figure under test conditions is given by:

$$\frac{signal + noise}{noise} = 20 \log \frac{V_1}{V_2}$$

$$\frac{signal + noise}{noise} = 20 \log \frac{1.0}{0.25} \tag{1}$$

$$signal + noise \cong 12\ dB$$

Additional reduction of the input signal should indicate observable detected signals with input levels of 0.1 μV or less.

5. Measure the $V_{FV}$ voltage required by the HFO varactor diode to produce the differential frequency

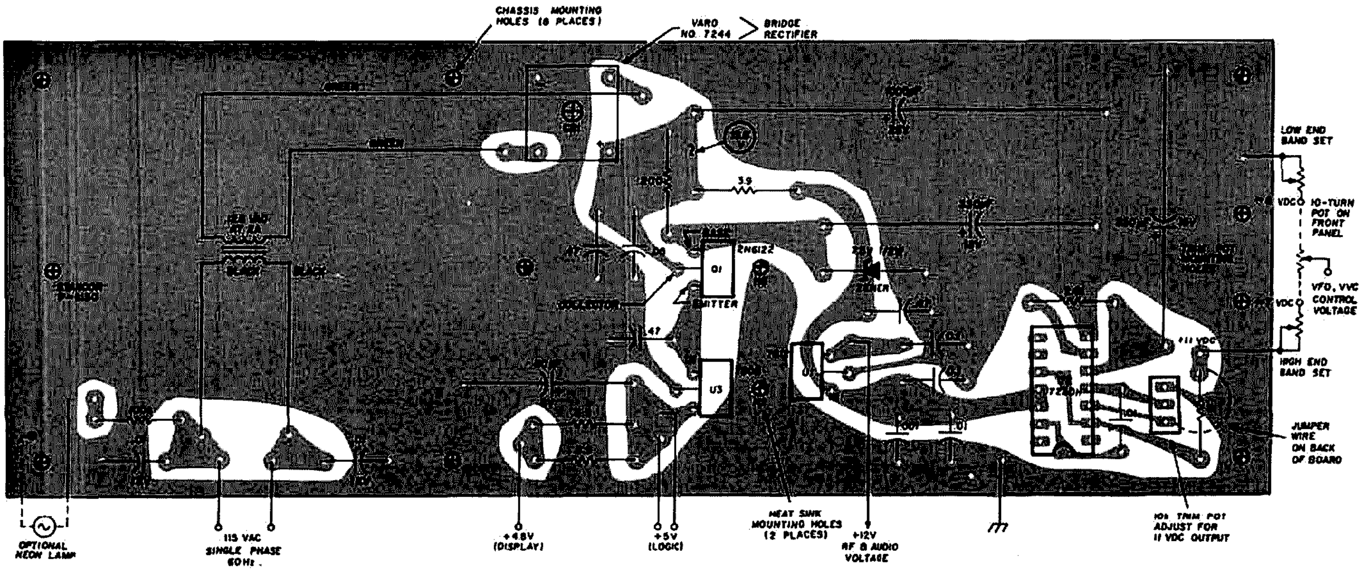



fig. 16. Power supply PC-board layout.

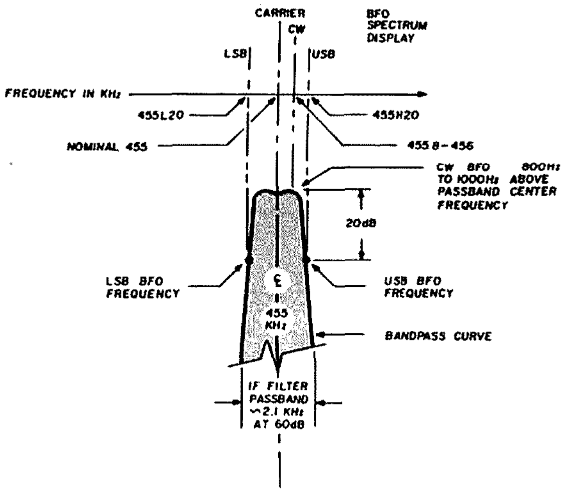



fig. 17. BFO versus i-f bandpass relationship (see text).

range desired; i.e., 12.5-12.85 MHz for standard 20-meter-band use. This voltage should be approximately 6-7.5 Vdc.

Apply ac line voltage to the power supply and adjust the trimpots for the $V_{FV}$ voltage to correspond to the measured values determined above.

**6.** Mount the power supply and interconnect the wiring. Using an external signal generator, introduce a 10-$\mu$V signal of 14.0 MHz and adjust the low end $V_{FV}$ voltage for detected audio. Do the same at 14.35 MHz, while adjusting the high-end trimpot on the power supply. Repeat as necessary for full-band coverage.

In part two of this article, a simple voltage-divider circuit is discussed, which will allow you to extend the tuning resolution of the HFO. At this point you should be familiar enough with the receiver operation to incorporate any special features desired. In part two, I will present the companion digital display circuit, and offer some useful suggestions.

## references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3. M. A. Chapman, K6SDX, "Varactor-Controlled Variable-Frequency Oscillator," *ham radio*, November, 1975, page 50.
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**ham radio**

# repeater kerchunk eliminator

## for mobile rigs

The ubiquitous 555 timer IC surfaces again — this time in a circuit that eliminates repeater noise in your receiver

Many repeaters, especially those on common channels, appear to be susceptible to endless testing and *kerchunking* by some mobile operators, usually to see if they are either getting into the machine, or if there is a repeater on a particular channel. Most of the time this is harmless, but quite often I've become irritated by so much noise and shut off the rig, thus missing an important call later.

The circuit shown here will eliminate most of these repeater squelch tails from just about any receiver employing its own squelch, but will allow normal communications to pass through. This is accomplished by placing the circuit between the receiver squelch gate and the point where the squelch acts on the audio amplifier. When a signal is received, and the receiver squelch gate is tripped, the kerchunk eliminator counts approximately 3 seconds before turning on the audio amplifier. If the received signal is gone before the 3 seconds is up, the radio remains silent, and the circuit resets itself. In this manner, the typical kerchunk will not be heard.

On the other hand, if the received signal lasts for longer than 3 seconds, which usually occurs if the repeater is interrogated, the receiver begins operating normally, squelch and all. Normal operation continues until the rig is turned off, then on again, which resets the kerchunk eliminator to give another 3-second squelch delay in the manner described.

The receiver also begins operating normally, without the 3-second squelch delay, if the operator uses the transmitter in the rig. In this way, a call may be made, and a reply received, without losing part of the message.

### circuit description

This circuit (**fig. 1**) is designed to work with receivers using a low voltage level to squelch the audio amplifier. If the input and output are inverted using switching stages similar to Q1, the circuit will then work with receivers using a high voltage level to squelch.

The NE-555 timer is wired for use as a monostable multivibrator (one-shot), and is triggered through inverting transistor Q1 and capacitor C1 each time the receiver squelch is tripped, provided that the scr is in the *off* mode. The output of the NE-555 goes

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high for about 3 seconds, which is determined by the R1, C2 time constant, and is applied to Q2 base. Transistor Q2, which now acts as the squelch gate, also receives the output of inverting transistor Q1, and remains saturated as long as Q1 is cut off (no incoming received signal) and/or the output of the NE-555 is high (3-second timer delay is activated). When Q2 is in the *cutoff* state, the receiver audio amplifier is turned on, and incoming signals are heard. In addition, the scr is triggered, which eliminates the NE-555 from the circuit, so the squelch may now operate normally through Q1 and Q2. This same result is also accomplished whenever the transmit power bus is activated for any length of time. The entire system is reset when power is temporarily removed from the rig.

You may have noticed by now that the NE-555 is grounded through a diode. This is required because the forward voltage drop of the triggered scr is too high (about 0.7 volt) to reset the NE-555. By floating the ground of the NE-555 slightly above that of the scr cathode, this problem is eliminated.

## construction and installation

The printed circuit board diagram and parts placement are shown in fig. 2. Before beginning construction, make sure that your rig has a place to mount a circuit board of this size and shape. Parts placement is not critical, of course. Mount the PC board some-

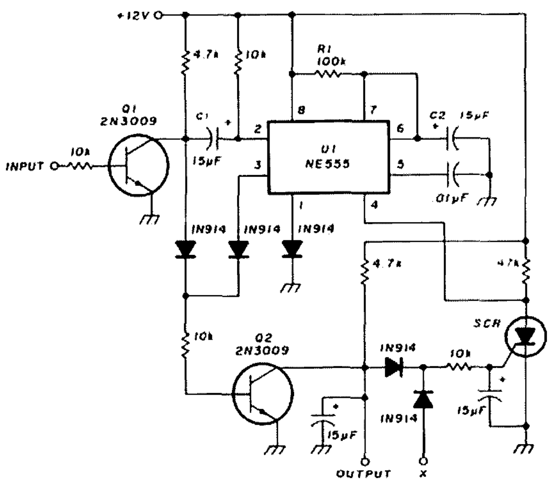


**fig. 1. Schematic of the repeater kerchunk eliminator. If the printed circuit shown in fig. 2 is used, resistors should be 1/4 watt, and the 15 μF capacitors either tantalums or vertical-mount electrolytics. Although a Radio Shack part number is given for the SCR, just about any type that will fit on the PC board will work. Likewise, any npn silicon switching transistor will work for Q1 and Q2. The input is routed from the squelch gate in the receiver, and the output connects to the squelch control point on the receiver audio amplifier.**

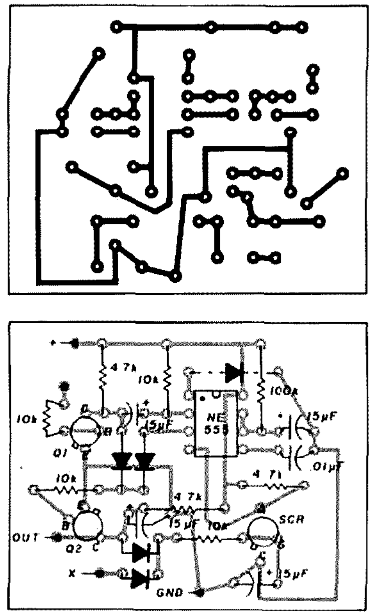

**fig. 2. Full-size printed circuit board (foil side) and parts-layout diagrams. The diode which grounds the NE-555 is mounted on the foil side of the PC board.**

where in the rig, and connect wires from $V_{cc}$, ground, and the transmit power bus from the TR switch to their appropriate places on the circuit board.

Now find the place where the squelch gate controls the audio amplifier in the rig, break the connection here, and run wires from these two points to the circuit board.

## testing

The system may be tested without an incoming signal by turning the squelch control fully counterclockwise. The characteristic background noise should become audible after about 3 seconds have elapsed. Turn the rig off, then on again, to reset the system with the squelch control set to eliminate background noise. Transmit briefly, and turn off the squelch again. Background noise should be audible with no delay encountered.

## results

After installing this kerchunk eliminator, I find that my rig is used much more often than before. An added advantage was discovered the next time I took a long trip: Distant repeaters that drifted noisily in and out of squelch did not activate the kerchunk eliminator until I was close enough to get a reasonably good signal. The kerchunk eliminator should prove to be a valuable addition to any mobile rig.

**ham radio**

# low-cost all-mode-protected power supply

Here's a power supply that is ideal for vhf transceivers and other equipment using solid-state devices

It seems that vhf repeater activity is the only game in town nowadays. However, with the many transceivers in use, many of which were probably purchased for mobile work, many amateurs have opted to use the rig in the house powered by ac line voltage. Now comes the hitch: a *good* power supply at reasonable cost.

The power supply described here, which I call the "gutless wonder," surely answers this need. Take a look at the schematic and you'll understand why I call it the gutless wonder: parts count is very low, but specifications are very high. Output is rock steady at 13.7 $\pm$0.7 volts dc using line voltage between 98 and 128 Vac. Regulation is within tenths of a volt, from zero to 5 amperes. The design includes short-circuit, overcurrent, and overvoltage protection. Because of the small parts count, a PC board isn't needed. A couple of small 6-lug terminal strips do the job for mounting most components, and a 4-lug strip is used for ac-line wiring. Simple, but effective!

## regulator

The heart of the supply is one of the 3-terminal regulator ICs currently available. Not long ago, I spent many hours working out empirical formulas to determine a circuit that would most efficiently provide proper output voltage with protective circuitry. Invariably problems occurred with tolerances and temperatures, which made trimming controls mandatory. Today, all you have to do is choose a voltage and current and, lo and behold! You can buy a small piece of plastic containing three leads that will do the job — the overall size is about 1 inch by 0.4 inch (25 by 10mm) including leads.

Included in this little device are overcurrent and thermal-protection circuits as well as tight regulation characteristics. The family is known as the 7800 series IC, made by Motorola, Fairchild, Signetics, and others. Many variations are available based on fixed output voltage, maximum allowable current, and packaging. Table 1 shows some of the devices

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**table 1. Rundown of some regulator ICs presently available for use in the power supply.**

| regulator type | package | Maximum load (amperes) | load current regulation (percent) | $P_D$ (free air) watts |
|---|---|---|---|---|
| 78XXC | TO3 | 1.5 | 2 | 3.0 |
| 78XXC | TO220 | 1.5 | 2 | 2.0 |
| 78MXXC | TO39 | 0.5 | 2 | 1.0 |
| 78MXXC | TO220 | 0.5 | 2 | 1.0 |
| 78LXXC | TO92 | 0.1 | 2 | 0.7 |
| 78LXXC | TO39 | 0.1 | 2 | 0.8 |
| 77XXC | TO220 | 0.75 | 2 | 2.0 |
| 77XXC | TO39 | 0.75 | 2 | 1.0 |
| 209 | TO3 | 1.0 | 2 | 3.5 (Fixed 5V output) |
| 309 | TO3 | 1.0 | 2 | 3.0 (Fixed 5V output) |
| 723C | TO100 | 0.15 | 0.6 | 0.85 |
| 723C | TO116 | 0.15 | 0.6 | 1.0 |

standard output voltages: 5, 6, 8, 12, 15, 20, 24. The 7700 series has an additional 20 volts.

presently available. In most cases, the chip must be mounted on a heat sink. The basic circuit using such a chip is shown in **fig. 1**. What could be simpler?

## the power supply

I wanted a minimum of 5 amperes at 13.6 volts for my power supply. No regulator chip is available that will handle this requirement, therefore it was necessary to add an external pass transistor to handle the required current. Also, no chips were available that provide exactly 13.6 volts (the resolution of this problem is discussed later).

The type 109-series regulators, familiar to most amateurs, uses a series-pass npn transistor to increase current-carrying capability. The circuit described here uses a pnp type MJ2955, which is the complement of the ubiquitous 2N3055. Any pnp transistor with a minimum $h_{fe}$ of 20 or more at 4 or 5 amperes, and a rated dissipation of more than 150 watts, will be adequate for Q2. The limiting factor will be temperature rise of the transistor when mounted in your heat sink.

The gutless wonder schematic is shown in **fig. 2**. No chips were available for the required voltage, so the regulator IC was put on a voltage pedestal, which is the secret of success in this supply. I used an 8-volt regulator chip arbitrarily; alternative values of resistance for R3, R4 are given should you wish to

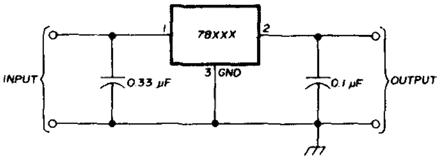



**fig. 1. Basic circuit of the three-terminal voltage-regulator.**

use a 12-volt regulator chip. Including all tolerances, the resulting output voltage was exactly as required.

The regulator chip specifications include a 5% tolerance. My output measured 7.8 volts, which is the average from the meter readings I used. If your supply output is off a couple millivolts, so be it.

## pedestal voltage

The pedestal voltage, $E_p$, is the voltage added to the rated chip voltage. It is approximately

$$E_p \approx I_p R4 + I_q R4$$

where

$$I_p \approx \left[\frac{E_o}{R3 + R4}\right]$$

$$E_o \approx E_{chip} + E_p$$

$$I_q = \text{typical chip quiescent current (4.3 mA)}.$$

Considering the interdependent parameters involved, it's best to use 40 mA for $I_p$. Operating thhe pedestal current at this value eliminated temperature and $I_q$ variations prevalent among the various regulator chips.

## circuit protection

As mentioned above, the 7800-series regulators are overcurrent protected, but the external pass transistor is not, hence the reason for adding one more pnp transistor, Q1, a 2N6049. This transistor should have a current rating of 2 amperes because, when turned on, it must carry the "short-circuit" current of the regulator chip, which is about 250 mA. R2 turns on Q1 at a specified current. R2 consists of ten 1.2-ohm 1/2-watt resistors in parallel to provide 0.12 ohm. When the load current reaches 5.4 amperes, Q1 turns on and its collector removes drive to Q2, causing the output voltage to decrease to zero.

Another feature of building the supply overcurrent protection in this manner is that, by adding another 1-ohm resistor in parallel with R2, the current limit can be increased to 6 amperes. Adding two 1.5-ohm resistors instead will increase the limit to 6.25 amperes; adding two 1-ohm resistors to R2 increases the current limit to 6.7 amperes. If you must go above 8 amperes, however, change the regulator chip to a 7808.

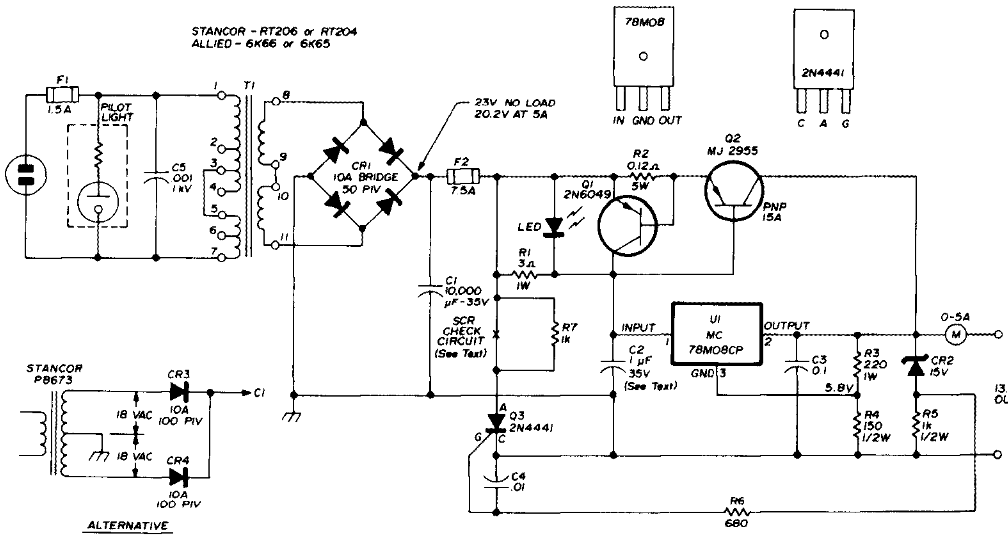
## heat sink

Transistor Q1 is a 40-watt device in a TO66 case, with an $h_{fe}$ of 20 at 1.5 ampere. Transistors Q1 and Q2 are mounted on a heat sink. The heat sink specified is ideal for this application, because the mounting holes are predrilled for both transistors. The bad news is that the heat sink isn't anodized; therefore both devices *must* be insulated from the heat sink. Be sure to use oxide-filled silicone grease (type DC 340) on *both* sides of the mica insulator. Use shoulder washers to insulate the transistor-mounting screws. Remember that the transistor cases are above ground, so avoid contact with external objects that may cause them to short circuit to ground.

## operating mode

This power supply operates in the so-called constant-voltage, current-limited mode. Let's be sure we understand what these fancy words mean.

The current through the pass transistor is limited by the value of R2; however, when the load is a low value of resistance or dead short circuit, a path to ground through the transistor causes the current to continue to flow, although it is limited. A voltage across Q2 exists, depending on the transformer secondary voltage, the electrolytic capacitor value, the transformer secondary resistance, and the impedance of the diodes in the bridge. In my supply this voltage turned out to be 16 volts. Multiply this voltage by the current, which is 5.4 amperes, and you'll find that the transistor is dissipating close to 90




| | |
|---|---|
| C1 | 10k μF, 35V electrolytic |
| C2 | 1 μF, 35V tantalum |
| C3 | 0.1 μF discap |
| C4 | 0.01 μF discap |
| C5 | 0.001 μF, 2kV discap |
| CR1 | bridge rectifier, 10A at 50 PIV |
| CR2 | 15V, 5% Zener |
| CR3,CR4 | 10A, 100 PIV |
| R1 | 3 ohms, 1 watt |
| R2 | 0.12 ohm, 5 watt (ten 1.2-ohm 1/2 watt resistors in parallel) |
| R3 | 220 ohms, 1 watt (alternative for 12V chip, 300 ohms) |
| R4 | 150 ohms, 1/2 watt (alternative for 12V chip, 39 ohms) |
| R5 | 1k, 1/2 watt |
| R6 | 680 ohms, 1/2 watt |
| R7 | 1k, 1/2 watt |
| T1 | multitap transformer. (For 6 amperes, Allied 6K66 or Stancor RT206; for 4 amperes, Allied 6K65 or Stancor RT204) |
| U1 | MC78M08CP (8V); 78M12 (12V) |

**Notes:**

1. Q1, Q2 are mounted on Radio Shack part no. 276-1361 heat sink. Mounting hardware: Radio Shack 276-1370, 276-1371.
2. U1, Q3 must be insulated from chassis.
3. Open jumper across R7 for checking crowbar circuit (see text).
4. LED short-circuit indicator must be a gallium-arsenide phosphide (GAsP) device; Vf approximately 1.8 volts.
5. T1 can be 18 Vac at 4 amperes for full-wave bridge rectifier. For a full-wave center-tapped circuit use Stancor P8673 or equivalent (see alternative circuit).

**fig. 2. Power-supply schematic. The voltage-pedestal method is used to obtain desired output voltage.**

watts! You'd better believe that the transistor gets *damn hot*, so the addition of the LED overcurrent indicator (described below) becomes more meaningful. If you need a higher capacity, at the very least mount the power devices on a *larger* heat sink!

## transient response

C2, C3 improve chip transient response. I've also found that C2 is a must to prevent oscillation under certain conditions. C2 can be anything from 0.33 $\mu$F and higher. Tantalum capacitors are recommended for their high-frequency bypass characteristics. If you can't find a tantalum capacitor, use a minimum of a 10-$\mu$F electrolytic with a 0.01-$\mu$F discap in parallel.

## overvoltage protection

A means of cutting off the power supply if an overvoltage condition occurs is the most important protective device I could add. Some transceivers have this protection built in — others don't. The thought of zapping a string of hard-to-find transistors in the transceiver for lack of a simple protective circuit is scary. The overvoltage-protection circuit shown is simple but effective — I know, because I tried it under dynamic operating conditions, and both primary and secondary fuses blew when 18 volts was applied to the meter end of zener CR2. The scr, Q3, is an 8-ampere rms device and I saw it do a job on the fuses! The scr turn-on time is 1 microsecond.

Capacitor C1 should be 10k $\mu$F at 35 volts. It can be made up of several caps in parallel if necessary. I've used capacitance as low as 3k $\mu$F, but the ripple voltage increased to a value higher than I cared to under high-current load conditions.

Transformer T1 was obtained from Allied Radio. A transformer with a secondary voltage of about 18 V at rated current is ideal but hard to come by. We must pay dearly for the multiple taps on the transformer, but I have the utmost confidence in the scrounging abilities of my fellow hams. Because of the intermittent nature of transmitter operation, a 3- or 4-ampere current rating should suffice for T1.

## construction

Photographs show the method of construction. Most of the wiring is no. 18 (1mm) wire to eliminate voltage drops. A separate terminal strip is used to mount the ten 1/2-watt resistors. As mentioned earlier, only two 6-lug terminal strips are required for mounting all the other components. A separate 4-lug strip is used for the 117-V wiring.

Both the regulator chip and the crowbar circuit scr are mounted on the bottom of the cabinet. Both must be insulated from the chassis using the mica washer supplied with the device. Use silicone grease here also. The regulator chip will handle up to 500 mA; however, above 200 mA the MJ2955 handles the output current. The MJ2955 and 2N6049 are both mounted on, but isolated from the heat sink with the wiring between the two made at the heat sink. Only four wires are needed from the terminal strip on the back cover through a hole in the cover to the power transistors. If you'd like to add a luxurious touch to your supply, install a 25 Vdc volt-meter and a 5-10 amp ammeter.

## checkout

Testing is done before connecting the supply to your rig. Turn on the supply and check the output voltage. If it doesn't come out right, note the voltages indicated in **fig. 2**. With so few parts involved, problems are usually caused by defective transistors. If you have power resistors, hook them up and check regulation. The output voltage shouldn't drop until you draw somewhat higher than 5 amperes, depending on the value of R1.

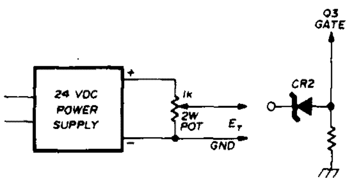



**fig. 3. Test setup to determine exact trip point of the SCR crowbar circuit in fig. 2.**

If you wish to check out the crowbar-circuit operation for the exact trip point, set up an external supply as shown in **fig. 3**. Disconnect CR2 from the output bus and remove the jumper from across R7. Connect a voltmeter between the scr anode and cathode. Turn on the supplies and rotate the arm of the pot toward the high-voltage end of the pot. Watch the voltage across the scr. When the test voltage fires the scr, the voltage across the scr will drop to about 1 volt. Measure the voltage between the pot arm and ground ($E_T$); this is the trip voltage. Remember to reconnect the zener and reinstall the jumper across R7.

The LED overcurrent indicator is turned on by the voltage across R1. It serves as a warning that something is wrong. If it lights up, turn off the supply and check the equipment connected as a load. Be sure you are using a GAsP LED — not a GaP; the voltage drops across the two are different. Fairchild, Monsanto, Litronix, and Hewlett-Packard devices are GAsP; Opcoa LEDs are GaP.

So now you have it. An excellent piece of equipment that will perform flawlessly under all conditions without jeopardizing your expensive transceiver.

**ham radio**

MICROPROCESSORS

# data converters: the microprocessor and the amateur

**In the future,** the microprocessor will be faced with many new and unique amateur radio applications. Unfortunately, though, a large problem exists when trying to interface the digital world of the microprocessor to the analog world of amateur radio. Of the numerous articles written on the microprocessor, many assume interfacing a foregone conclusion, being contained within the microprocessor. Nothing could be further from the truth! This article will provide the necessary guidance to fill this void by demonstrating how data converters perform this valuable interface.

Two types of data converters will be discussed; one changes voltage to a digital code and the other performs the opposite function, transforming a digital code to an analog voltage or current. The former is an analog-to-digital converter (ADC), while the latter is a digital-to-analog (DAC). The digital code is in the format necessary for processing by the microprocessor. In effect, the two devices provide a means by which the microprocessor can acquire analog information, massage it, and deliver an output, in analog form, to an external source or terminal capable of controlling or displaying a function of the original analog input.

It may appear that a number of circuits or black boxes other than a microprocessor can do this. Certainly, but remember, this type of circuit is a fixed design, its cost is high, and extremely costly if hardware changes are required to change functions. The microprocessor/data converter combination is superior due to its low initial cost (by comparison) and the flexibility of being able to handle a multitude of assignments with no change in hardware. The microprocessor/data converter system is used in industry to control such variables as temperature, flow, pH, viscosity, and weight of a liquified processed chemical. Furthermore, if the process formula is ever changed the hardware is left intact while the program is altered. It is this feature that makes the microprocessor/data converter system attractive.

Choosing the correct ADC or DAC is based to a great extent on the type of analog input and output to be associated with the respective data converters. There are many types of data converters that provide an assortment of functions, but those most applicable to amateur radio are the integrating and successive approximation ADC, and the voltage/current and multiplying DAC. A description of the functional peculiarities of each type is necessary to fully understand their application.

## digital-to-analog converter

Fig. 1 illustrates a simple *n-bit* binary DAC. The digital input is applied to the terminals labeled LSB to MSB. Each digital input drives an fet switch that connects the binary weighted summing resistors to the reference or to ground depending on the individual state of each of the digital inputs. Note that by virtue of the R-2R resistor configuration, binary weighting is achieved as the current flowing through each resistor is divided by a factor of two at each junction. With this scheme, the impedance as seen by the op

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amp is constant and only two resistor values are required.

The output voltage ($E_0$) is directly related to the number of switches which are connected to the reference by the 1 applied to the digital input. On the other hand, a zero applied to the digital input grounds the R-2R resistors, resulting in zero voltage at the output of the DAC. Thus, all 1s would give a maximum voltage at the $E_0$ and all 0s, of course, is a binary code which produces an output voltage directly proportional to the digital code. This type of DAC delivers output voltage by virtue of the op amp inserted at the summing point and will, in most cases, fill the need of most amateurs. A current DAC does not contain an output op amp and delivers a current to its load.

The current DAC, due to its fast settling time and small glitches, is ideal when the interface is with a CRT display. The glitches are voltage spikes on the output that are caused by the switching transients. They are particularly large at the major transitions, i.e., going from all 1s to all 0s and vice versa. One method of reducing the spikes is to insert a lowpass filter before the display; another is to use a deglitcher after a current DAC. This is essentially a sample-and-hold type circuit, usually built from discrete components.

There may be times when it's necessary to multiply a digital word by a range of analog voltages to pro-

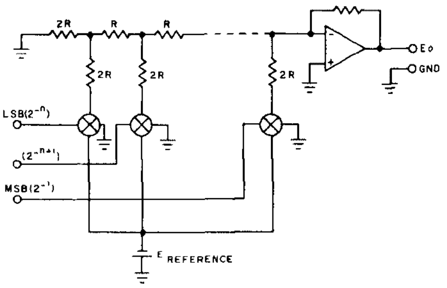


**fig. 1. Schematic diagram of a basic digital-to-analog converter. The fet switches connect the 2R resistors to either ground or the reference voltage.**

duce an output voltage or current representing their product. This is accomplished through the use of a multiplying DAC (MDAC) in which the internal reference is supplemented by an analog input. Of course, it's possible to use the MDAC as a conventional voltage or current DAC by applying an external reference voltage to the analog input terminals. This becomes particularly advantageous when more than one DAC is used in a system. A number of MDACs can share the same reference source, thus saving money on the cost of a conventional DAC and at the same time making the stability characteristic of each MDAC identical to one another by virtue of the common reference voltage.

## analog-to-digital converter

The successive approximation ADC is shown in **fig. 2** and functions as follows. First, the most significant bit (MSB) of the DAC is turned on by the control logic, and the output is compared to the ana-

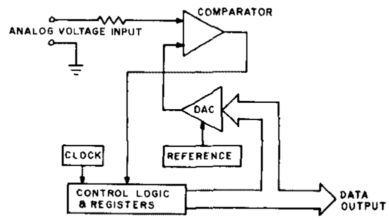


**fig. 2. Block diagram of a successive approximation type analog-to-digital converter. The analog voltage is compared to an internally generated voltage. The control circuitry that operates the internal DAC also generates the final digital output.**

log input voltage. If the comparator determines that the input is less than the DAC output, the bit (MSB) is turned off. If it is greater than the DAC output, it is left on.

The next MSB is then turned on and the same sequence of events repeated adding to the previous bits whether left on or turned off. This process is continued until a digital code (binary, 2's complement, offset binary, etc.) is formed, producing a voltage output from the DAC that in turn produces zero output from the comparator. In short, the output voltage from the DAC equals the analog input voltage to the ADC.

The state of the code present at the output registers is now a digital representation of the input analog voltage. Actually, this digital output does not cover each finite voltage applied to the input but a quantized incremental voltage step, the number of which is determined by the resolution of the ADC. An ADC with a resolution of 12 bits, for example, would have 4096 discrete quantized voltage levels or digital states. A 10-bit unit would have 1024, an 8-bit, 256.

The integrating ADC has a much slower conversion speed than the successive approximation type, but is a less complicated device and therefore, less expensive. It is used primarily with a slowly changing analog input as from a thermistor, thermocouple, or strain gauge input. An additional advantage of most integrating ADCs is their ability to convert in 16.67 microseconds, giving them exceedingly high common mode rejection to 60 Hz.

This type ADC, as shown in **fig. 3** , integrates the

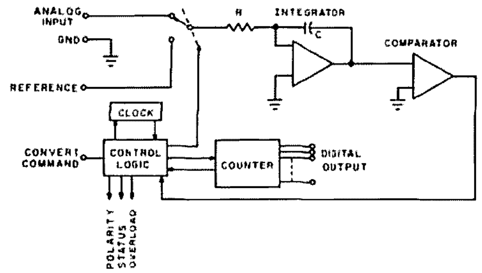


**fig. 3. An integrator type ADC is used for slowly varying analog inputs.**

analog input signal for a fixed period of time determined by a counter. The resulting integrated voltage is returned to zero by integrating a reference voltage of the opposite polarity. The resultant time of the integrator to return to zero as measured by the counter is proportional to the average value of the analog input over the integration period. Pulses from the counter are loaded into a register and deliver a specific output code to the microprocessor. ADCs of this type frequently have other options such as polarity indication, over-range output, ratiometric measurement capability, and auto zero. They are the basic ingredients of a digital panel meter and the digital voltmeter.

## choosing the right converter

Speed is a major consideration; this is known as settling time for DACs and conversion time for ADCs. In the former case, the time it takes a DAC to reach its rated output voltage/current after strobing is of prime concern when considering the response of the circuit it feeds.

The DAC, of course, does not respond coincidentally with its command to convert. It takes time for the fet switches to close and the op amp to reach its proper output magnitude. ADC conversion time is selected as a function of how rapidly the input analog voltage varies. A general rule of thumb requires the conversion time to be at least twice that of the input frequency in order to digitally reproduce the input analog voltage.

If a thermal input representing rf power were applied to an ADC, for example, its slowly varying nature would dictate the use of an integrating ADC. On the other hand, digitizing voice frequencies would require the use of successive approximation type. Assuming the maximum voice frequency to be digitized is 5 kHz, a conversion time of no more than 100 microseconds would be required to produce an accurate digital representation of the input (a faster conversion speed would more precisely duplicate the input voltage).

When the rate of the signal being digitized also

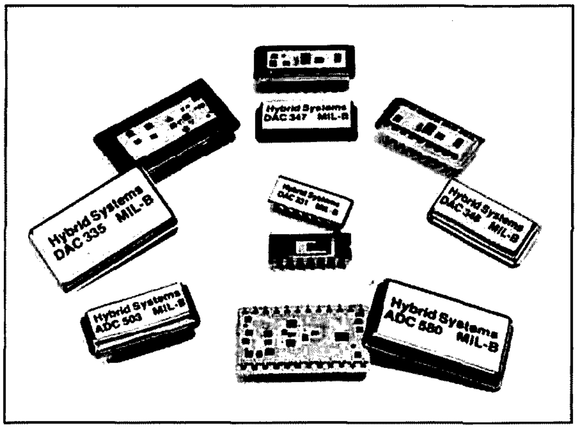


**table 1. Manufacturers of data converters.**

| | |
|---|---|
| Analog Devices, Inc.<br>P.O. Box 280<br>Norwood, Massachusetts 02062<br>(617) 329-4700 | Hybrid Systems Corp.<br>Crosby Drive<br>Bedford, Massachusetts 01730<br>(617) 275-1570 |
| Analogic Corp.<br>Audubon Road<br>Wakefield, Massachusetts 01880<br>(617) 246-0300 | ILC Data Devices Corp.<br>Airport Int'l Plaza<br>Bohemia, New York 11716<br>(516) 567-5600 |
| Beckman Instruments, Inc.<br>Helipot Division<br>2500 Harbor Blvd.<br>Fullerton, California 92634<br>(714) 871-4848 | Intech Function Modules<br>282 Brokaw Road<br>Santa Clara, California 95050<br>(408) 244-0500 |
| Burr Brown<br>Int'l Airport Industrial Park<br>P.O. Box 11400<br>Tuscon, Arizona 85734<br>(602) 294-1431 | Micronetworks Corp.<br>324 Clark Street<br>Worcester, Massachusetts 01606<br>(617) 852-5400 |
| Cycon, Inc.<br>1240 Elko Drive<br>Sunnyvale, California 94086<br>(408) 734-1535 | *Motorola Semiconductor Products*<br>Box 20912<br>Phoenix, Arizona 85036<br>Contact Distributors |
| Datel Systems, Inc.<br>1020 Turnpike Street<br>Canton, Massachusetts 02021<br>(617) 828-6395 | National Semiconductor Corp.<br>2900 Semiconductor Drive<br>Santa Clara, California 95051<br>Contact Distributors |
| Dynamic Measurements Corp.<br>6 Lowell Street<br>Winchester, Massachusetts 01890<br>(617) 729-7870 | Phoenix Data, Inc.<br>3384 West Osborn Road<br>Phoenix, Arizona 85017<br>(602) 278-8528 |
| Teledyne Philbrick<br>Allied Drive<br>Dedham, Massachusetts 02026<br>(617) 329-1600 | Precision Monolithics, Inc.<br>1500 Space Park Drive<br>Santa Clara, California 95050<br>(408) 246-9222 |
| Teledyne Semiconductor<br>1300 Terra Bella Avenue<br>Mt. View, California 94043<br>(415) 968-9241 | Zeltex, Inc.<br>940 Detroit Avenue<br>Concord, California 94518<br>(415) 686-6600 |


varies at a rapid rate, it becomes necessary to retain the original signal for a period of time sufficient to perform the digital conversion. Simply feeding this signal to the ADC will not guarantee an accurate conversion. However, this can readily be achieved by inserting a sample-and-hold circuit between the analog input and the input to the ADC. The sample-and-hold will track the input signal, acquire it on command, and hold it while the ADC performs the conversion. At the completion of conversion the ADC generates a reset pulse and puts the sample-and-hold back into the sample or track mode. This sequence is repeated successively as determined by the nature of the measurement or control function desired.

Resolution, or the number of bits, of either the DAC or ADC determines the incremental accuracy and the linearity. In effect, it determines the number of increments the analog input or output has been divided into. A high resolution converter naturally yields a more accurate representation of its input over the complete range of its input than would a lower resolution device.

There are many specifications that are related to the converter's temperature environment. Close examination of each manufacturer's specification is necessary to understand a converter's parameters when subject to changes in temperature. Reference is made to the bibliography at the end of this article for those readers who wish more detailed knowledge on data converters. Many of the manufacturers listed in **table 1** have applications notes and technical papers available which discuss the various details of data converter products.

## interfacing

The ADC and the DAC in **fig. 4** represent memory locations for the microprocessor similar to those occupied by RAMs and ROMs that are normally associated with microprocessor/microcomputer systems. Access to the converters is gained by the assignment of an address code similar to those allotted to conventional peripherals such as displays, *Teletype* machines, and printers. This arrangement provides direct access to the converters without the necessity of using the microprocessor memory.

Although **fig. 4** does not show it, a *programmable* or *peripheral* interface element can be inserted between the ADC and the microprocessor. This IC handles addressing and decoding generation of instructions, interrupt, and bus access signals.

Controlling a DAC is a simple process and is accomplished by performing a store operation; that is, transferring data from the microprocessor to the latches or registers associated with the DAC. The DAC now only requires an instruction to deliver its output to its load. The ADC is accessed by fetching

data from its memory location or by having the programmable or peripheral interface element perform this function under instructions from the microprocessor. Of course, if the total system involves a multichannel arrangement of DACs or multiplexed ADCs, each individual channel must be separately addressed and fetched. Systems this complex depend entirely on the programmable and peripheral interface element to perform the addressing.

This has been an overview of how the amateur can utilize the microprocessor to automatically control his equipment or perform measurements in a

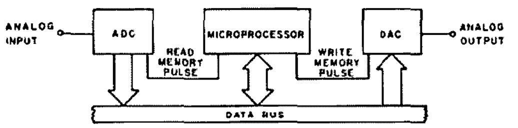


**fig. 4. Block diagram of the interface between a microprocessor and data converters. Either data converter can be eliminated, depending upon the analog voltage requirements.**

systematic fashion. It shows the necessity of the data converters as a means of connecting the microprocessor to a vco, a voltage-controlled attenuator, an rf power meter, or other device that performs equipment adjustments or measurements. It will hopefully provide many amateur radio enthusiasts with a tool that will provide many interesting hours of thought and experimentation.

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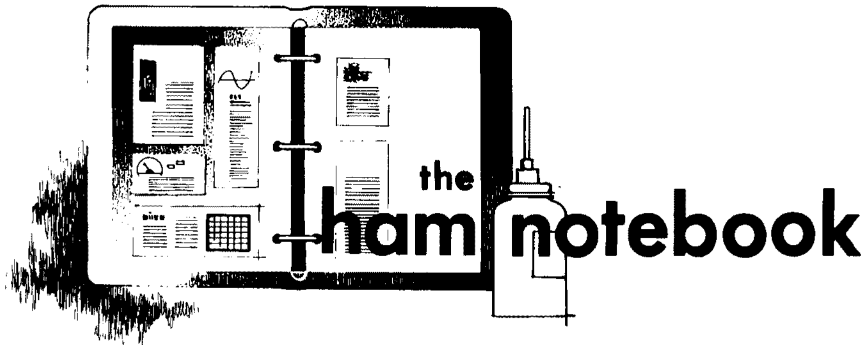
**ham radio**




## increasing the capacity of the RAM RTTY message generator

I was very interested in K4EEU's 146-character message generator (*ham radio*, January, 1975, page 8) and immediately built it. I am particularly impressed with its performance in conjunction with my ST-6. I recently changed the clock because the Motorola MC724P is now reserved for maintenance, and I used the double monostable 74123 which is more readily available.

From time-to-time the number of available characters was not sufficient, so after I discovered INTEL's 2102 (1024-bit RAM), I decided to multiply the original capacity of 146 characters by four. Now I can store 585 seven-bit RTTY words. The cell array of the 2102 is organized is 32 rows and 32 columns; that is to say, it's necessary to have two five-bit counters instead of the two four-bit counters (7493) for correctly decoding the 1024 positions.

It is possible to add one bit to each of the four-bit counters with the state of the fifth bit changing each time output D of the 7493 falls from logic 1 to logic 0. A dual JK master-slave

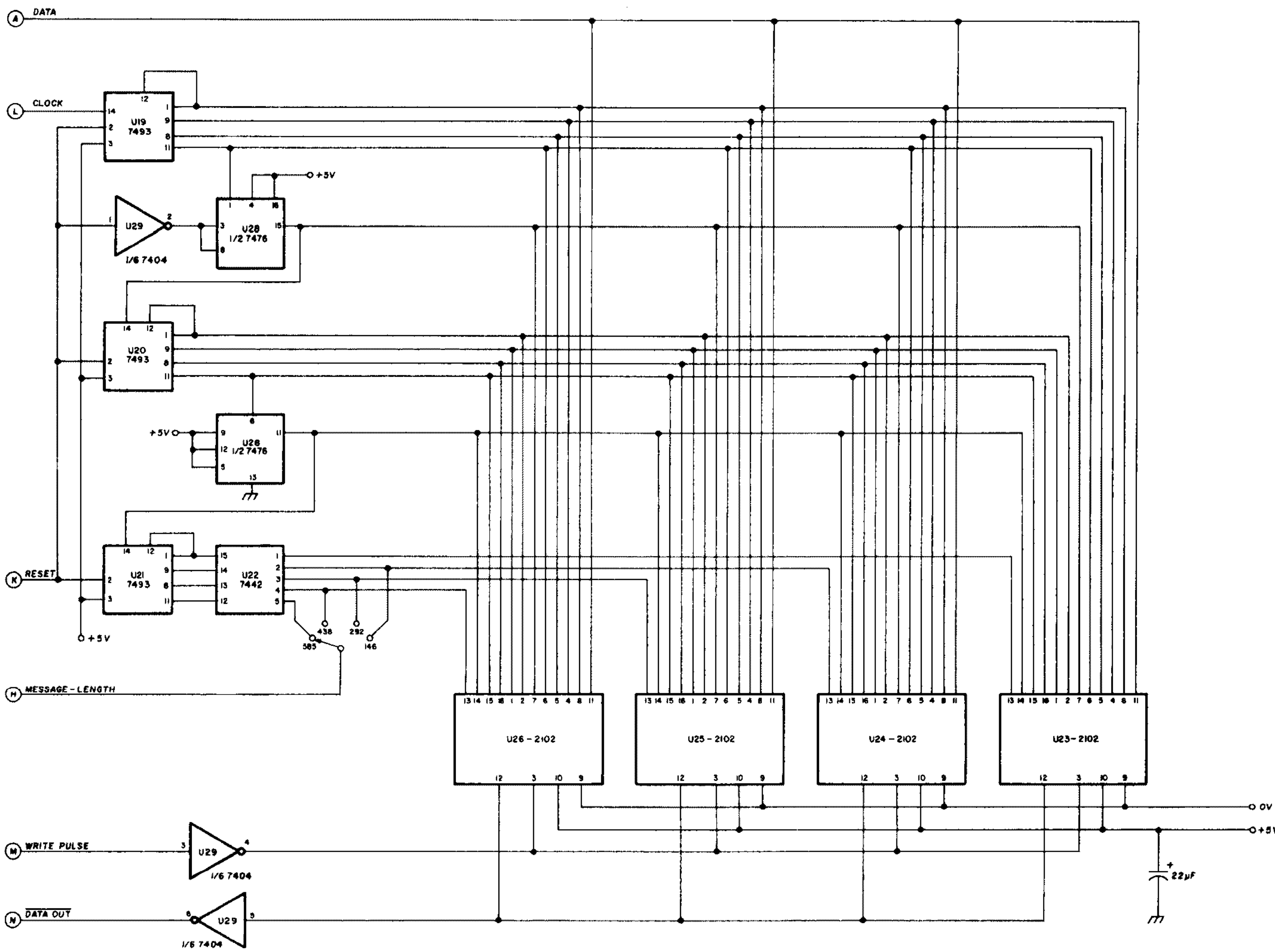



fig. 1. Schematic diagram of the improved RTTY message generator.

flip-flop (7476) does the job nicely by connecting one flip-flop to the D output of U19 and the other to the D output of U20.

To be sure that the Q output is set to zero when beginning a message, it is necessary to apply a zero to the clear input at the same time U19, U20 and U21 are reset. In these circuits the reset level is logic 1, so it is inverted by one-sixth of a 7404. With this arrangement, the 1024 positions of each 2102 memory can be decoded correctly, the chip-select circuit being the same.

In addition to the different pin configurations of the 1101 and 2102, two other points are changed. First, the read-write levels are inverted, so I used another part of the 7404 to correctly set the levels. Second, the 2102 offers only a DATA OUT and we must have a $\overline{\text{DATA OUT}}$, so I inverted the output of the memories with an inverter (part of the 7404). Everything is now set to enter 585 RTTY characters and recall them at will.

The new integrated circuits were mounted on wire-wrap sockets installed in a 0.1 by 0.1 inch (2.5 by 2.5 mm) grid-drilled epoxy board. The connections between these circuits have been wrapped, the 1101s removed from their sockets, and the connections between the original circuit and the additional circuit are hard-wired with 10-inch (25cm) wires.

The first test was okay and the entire system worked well with the new, added circuit. It is possible to install the new circuit on spacers in the former box, but it would be better to design a printed-circuit board that could be connected directly to the 1101 sockets. In the latter case, only seven connections would have to be made with single-contact connectors.

**Jean-Claude Piat, F2ES**

## Heath SB-102 headphone operation

The *phones* jack on the SB-102 is fed from a high-impedance source and should be used with headsets of high impedance. If low-impedance headphones are used, the *phone volume* control must be wide open to achieve a good balance between speaker and headsets. To provide a

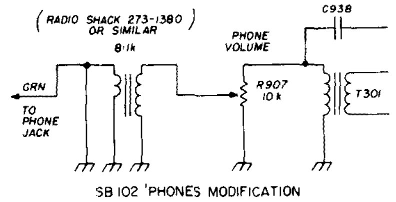



**fig. 2. Shows addition of a small output transformer to the headphone circuit of a Heathkit SB-102 to permit use of low-impedance headphones.**

better impedance match and control over the volume, a small imported output transformer was used as shown in **fig. 2**. Rather than add the circuitry externally, the transformer was mounted on the *bias* pot by soldering the frame to the pot body with the primary high-impedance winding facing the rear of the chassis and the *phone volume* pot. A two-terminal lug was attached to the chassis with one of the circuit-board mounting screws opposite the *bias* pot. The leads were transferred from the *phone volume* pot to the terminal strip, the secondary of the transformer wired to the terminal strip and the primary to the *phone volume* pot. Now, balance between speaker and headphones is more easily achieved and no distortion is noted.

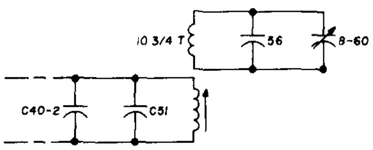
**Paul K. Pagel, K1KXA**

## NCX-500 modification for 15 meters

If the NCX-500 tranceiver is not carefully aligned, the fourth harmonic of its 5.2-MHz carrier oscillator, which appears at 20.8 MHz, creates a strong spurious signal in the receiver. A simple trap installed on the 15-meter mixer coil minimizes this spurious response.

The trap consists of a coil and parallel capacitor, prepared in the following manner.

Close wind 10-3/4 turns of no. 18 AWG (1.0mm) enameled wire on a 3/8-inch (9.5mm) diameter, 1-inch long (25mm) coil form. Place a 56 pF fixed capacitor in parallel with an 8-60 pF compression trimmer-type variable capacitor and place the parallel combination across the coil. Wrap approximately two turns of electrical tape around the top of L14 to obtain a tight fit and fit the trap to the top of L14 of the NCX-500. Adjust the coils so that the spacing between the two windings is about 5/8-inch (16mm).




### adjustment instructions

**1.** Remove crystal Y3 (25.1 MHz) from the transmitter.

**2.** Tune the transmitter to the 20.8 MHz spurious frequency.

**3.** Adjust the variable capacitor on the trap for **minimum** transmitter output or plate current reading. (Use whichever gives the most sensitive indication).

**4.** Re-adjust the distance between the windings by moving the trap a little closer or further away from coil L14 and retune the trap as per **step 3**.

**5.** Replace the 25.1 MHz crystal, retune the transmitter to 21.1 MHz and re-adjust the slug in L14 for **maximum** transmitter power output. Note that the core in L14 should be sitting at the top end of the coil, farthest from the chassis.

6. Repeat **steps 1** thru **5**, because adjustment of either the trap or coil L14 will influence the other.

The trap will effect the performance of the NCX-500 receiver section at 21 MHz, hence coupling between L14 and the trap should be as loose as possible to minimize this effect. Finally, fix the trap to coil L14, using a little adhesive such as *Duco* cement.

**Conrad J. Espinola WA1KYO**

# NEW products

## The French Atlantic Affair

In the midst of so much mistaken publicity about amateur radio, wherein problems caused by CB stations are blamed on the ham by an uninformed press, it is refreshing to see a book that has amateur radio woven into the plot as one of the "good guys." Moreover, the book is not one that would normally rate just a casual glance and then be placed back on the shelf.

The French Atlantic Affair, by Ernest Lehman, is a novel of such length that it will keep the average reader at it for several evenings, if he can force himself to put it down, or all night if he cannot. There is almost continuous action throughout the story, and the way author Lehman has brought amateur radio into focus early in the converging lives of the characters portrayed almost guarantees that it will play a significant part in the unfolding story.

Not that hams are all portrayed as knights in shining armor — their feet of clay are clearly evident so that the uninformed reader can see that they are ordinary human beings like everyone else. As such they have their hang-ups, problems, and families whose patience with this obsession wears thin at times. The attitudes and opinions about hams and their technical vices are all too familiar to many of us who have been the focal point of similar comments from friends and wives who are not completely dedicated to the hobby of chasing DX or burning the midnight hours away conversing with a stranger in a far corner of the land.

To an experienced amateur, it would seem that the author knows about ham radio from the inside — the equipment is real, the procedures are correct, the language is right, and the bands go dead at the right time; there is even the right kind of nit-wit who clobbers the frequency in the middle of an important exchange. To balance the scale against the baddies, there is the good guy who provides a phone-patch across the continent to overcome the dead-band problem, and a very understanding ham on Long Island who cancels out of an important golf engagement to take part in the high drama.

The author *does* know about amateur radio — he is K6DXK, and has been as busy on the air as at the typewriter. The way he tells it, the time spent at the ham rig at times overshadowed the writing, so the book was somewhat of a combination of the two important factors in his life.

However, Lehman is definitely accustomed to writing top-grade material; he has written screen plays for several well-known films such as *West Side Story, The King and I*, and *Who's Afraid of Virginia Woolf*, and others. He has received six Writer's Guild awards and six Academy Award nominations.

*The French Atlantic Affair* was not written for hams alone — you don't pin your hopes of making a best seller on a potential audience of a few hundred thousand. The book is written to appeal to people who would be spellbound with such stories as *Hotel, Airport*, and *Where Eagles Dare*; it can be very favorable compared to these best sellers. The action is almost nonstop, and one segment builds suspense that is guaranteed to evoke interest in the next page or chapter — this reviewer suffered the consequences of lack of sleep all the next day because of the irrestible urge to see how things ended.

The book is recommended for adult reading, and, while it is unlikely that any of you hams who read it will become heroes of the level attained by the operators involved in this story, you will be comforted to know that you are not alone in being a member of a group whose hobby, ability, and dedication is misunderstood and underrated.

Amateur radio plays a vital part in *The French Atlantic Affair*, and the resulting publicity should be most beneficial to our hobby and image. If the story makes it into a hit movie (the screen rights have been purchased by MGM), that's all to the better.

*The French Atlantic Affair* has been selected by the Literary Guild and by the Playboy Book Club, and is being published by Atheneum Publishers; it is available from Ham Radio's Communications Bookstore, Greenville, New Hampshire 03048; $10.95 postpaid. Not recommended for young readers.


More Details? CHECK—OFF Page 126

## 80-Meter DX handbook

To a large number of amateurs, the 80- or 75-meter band is most useful as a rag-chewing spot. Daytime coverage is strictly local, and long paths open during the dark hours. Average contacts take place across distances that reach from one coast to the other during favorable conditions.

However, in the true amateur spirit, there is a core of DXers who are constantly probing to see what can be done on this "local" band to extend the communications range to the farthest parts of the globe.

John Devoldere, ON4UN, has compiled a handbook of things to do to squeeze the most out of this rag-chewer's haven; the *80-Meter DX Handbook*.

The handbook is a complete works on the art of DXing in this part of the HF spectrum, from why it happens (or does not happen), to what to build and how to use it. The book is arranged in four chapters: Propagation, Antennas, Stations, and Operating Practices. The propagation section covers magnetic disturbances, seasonal effects, twilight periods, paths, and more. As might be expected, the antenna section is the largest part of the book, with 15 distinct subjects — starting with the fundamentals and working through ZL-specials and Beverage antennas.

The section on stations contains a concise discussion of popular transmitters and receivers, and some hints as to their proper use. In Operating Practices, the author gives you information about what band segments are used in various countries, talks about procedures, types of operation, and points out some of the awards that have been won, and some that are much sought after.

The *80-Meter DX Handbook* ends up with a very good Bibliography for those who would like to explore some of the reference works in greater detail.

Published by Communications Technology, Inc., the *80-Meter DX*

*Handbook*, by ON4UN, is available from Ham Radio's Communications Bookstore, Greenville, New Hampshire 03048. Order HR-80M, $4.50 postage paid.

## active filter design

This new book by Claude Lindquist is an invaluable reference for engineers, technicians, and amateurs who must understand signal processing and filtering, and who must specify, design, or adapt active filters to their own uses. The book features standard design curves and tables, practical design examples such as *Touch-Tone*, Dolby B, amplitude and delay equalizers, and presents numerous problems with solutions (which form an encyclopedia of active filter design). Also included are extensive references and unpublished research results (including delay analysis, high-frequency filters, tuned filters, discriminators, etc.).

Hardbound, 749 pages, $21.95 postpaid from Steward & Sons, Post Office Box 15282, Long Beach, California 90815.

## printed circuit board kit

Excel Circuits Company has recently introduced a kit for making your own printed circuit boards. The kit contains materials that will allow you to produce your own boards, starting with the artwork and ending with a finished board. In contrast to other processes, this kit uses a photo-positive method, meaning that no image reversal and hence darkroom facilities are required.

Included in the materials are sufficient supplies to do the layout of small PC boards. The kit contains integrated circuit layouts, tape and pads, 10 x 10 graph paper, and a mylar-based paper to hold the layout. With no image reversal being necessary, the finished layout can be used to expose the board. One other advantage of using a photo-positive system is that dirt and dust on the

layout will not cause holes in the eventual circuit traces. Also, the neat plastic container can be used to hold the etchant while etching a board.

From start to finish, the Excel Circuits Company has provided materials that permit the fabrication of quality circuit boards. In addition, they've eliminated the problems of obtaining supplies from different sources. The price of the complete kit is $19.95, postpaid. For more information write to Excel Circuits Company, 4412 Fernlee, Royal Oak, Michigan 48073.

## new wire-wrapping tool does all


The new WSU-30M "Hobby Wrap" tool performs the complete wire-wrapping function. First, the tool wraps 30 AWG (0.25mm) wire onto standard 0.025-inch (0.6mm) square DIP socket posts. Also, the tool unwraps and, finally strips 30 AWG (0.25mm) wire nick-free through a unique stripping blade.

Particularly important, the WSU-30M produces the "modified" style of wrap, in which approximately 1½ turns of insulated wire are wrapped in addition to the bare wire for added mechanical stability.

Designed for the serious amateur, the WSU-30M features compact, all-metal construction for years of dependable service. At only $6.95, this tool is a remarkable value, performing the work of three separate tools at a fraction of their comparable prices. It's available from your local electronics distributor or directly from O.K. Machine and Tool Corporation, 3455 Conner Street, Bronx, New York 10475.

$1.50

# ham radio magazine

## NOVEMBER 1977



## third annual receiver issue

# ham radio magazine

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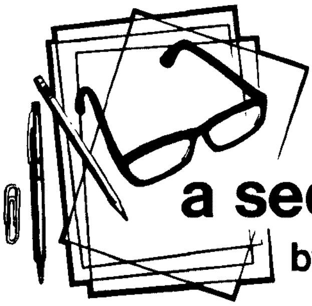


# a second look

by Jim Fisk

**It was just thirty years ago**, on December 23rd, 1947, to be exact, that a group of scientists at Bell Laboratories built a one-stage amplifier circuit around the world's first transistor, giving birth to a whole new era of electronics and communications. But the beginning of the story was not in 1947, but long before. There had been hints of amplification in semiconductors as early as the 1920s but few experimenters could duplicate the results. Nobody realized the effect of semiconductor impurities nor understood the action of semiconductor materials.

In 1930, Dr. Julius Lilienfield, a German physicist, actually patented a semiconductor amplifier that could be compared to today's mosfet. Although Dr. Lilienfield's amplifier worked, it could not be duplicated by other workers, and it slowly slipped into oblivion.

In 1939, Dr. William Shockley made an entry into his lab notebook at Bell Labs, "It has today occurred to me that an amplifier using semiconductors rather than vacuum is in principle possible." It was nearly eight years before this concept would bear fruit. A large part of this period was spent in learning more about that old bugaboo, semiconductor impurities.

The 1N21 crystal detector, developed during World War II and the workhorse of wartime radar receivers, provided some of the impetus. After the war a solid-state research team at Bell Labs, co-headed by Dr. Schockley, started experimenting with germanium and silicon, two semiconductors that were easy to work with. As one of the group said recently, "We felt that the area was so fertile that you could devise an experiment in the morning, go out in the lab and try it in the afternoon, and write a paper about it that evening."

The first device the group attempted to build was what is now called an insulated-gate fet. The device didn't work. The group scrambled around, dug into the literature, and spent long hours discussing the alternatives.

Dr. Walter Brattain tried an experiment where he covered a metal point with a thin layer of wax and pushed it down on the surface of a piece of silicon. He then surrounded the point with a drop of water and made contact to it. The water was insulated from the point by the wax layer. He found that voltages applied between the water and the silicon would change the current flowing from the silicon to the point. Power amplification had been achieved! Unfortunately, the drop of water evaporated almost as soon as things were working well.

This led to experiments with other electrolytes that didn't evaporate so readily. Then, they discovered a thin oxide layer on the surface of the semiconductor under the electrolyte and decided to eliminate the electrolyte and use a spot of gold as a field electrode.

When this was tried, an electrical discharge between the point and the gold spoiled a spot in the middle — when they washed off the electrolyte they had inadvertently washed off the oxide film, which was soluble in water. However, by placing the point around the edge of the gold spot they observed a new effect — when a small positive voltage was applied to the gold, the current flow was greatly increased. Four days later two gold contacts less than two-thousandths of an inch apart were made to the same piece of germanium and the first transistor was born.

Nine years later, in 1956, the three inventors, Dr. William Shockley, Dr. John Bardeen, and Dr. Walter Brattain were awarded the Nobel prize in physics. Little did they realize that their crude laboratory device would spawn a multi-billion dollar semiconductor industry that today affects all our lives.

**Jim Fisk, W1HR**
**editor-in-chief**

# presstop

SEPARATE REPEATER LICENSES will no longer be required and an Amateur operating a repeater will be able to do so simply by signing his regular call with the suffix "/RPT" on CW or "Repeater" on phone when in the repeat mode. This far-reaching deregulatory action came as a Report and Order on Docket 21033. Though agreement was universal throughout the Commission that separate repeater licenses were useful, it was agreed they required more Commission investment in time and money than they were worth.

The Repeater Subbands were also expanded, with an additional one MHz — 144.5 to 145.5 — now opened to repeaters on two meters and all Amateur frequencies above 220 MHz, with the exception of the 435-438 MHz space communications slot, now available for repeater use. The 10- and 6-meter repeater limits remain unchanged.

Technicians Will Receive another 500 kHz, down to 144.5 MHz, so they'll be able to use the new two-meter repeater subband, which neatly avoids the OSCAR activity just below 146 and also manages to straddle (assuming 600 kHz input-output separation) existing SSB and A-M simplex operation between 145.0 and 145.2. An interesting suggestion from W3LOY is that users of the new repeater subband standardize on 20 kHz channels, providing more repeater pairs than 30 kHz but without the adjacent channel problems that have plagued many users of the 15-kHz split.

20-kHz Channel Spacing for the new 2-meter repeater subband is receiving very strong support nationwide. The Northern Amateur Relay Council met in Sacramento the end of September and unanimously proposed a band plan with 144.51-144.89 for repeater inputs and 145.11-145.49 for corresponding repeater outputs. They designated the 144.9-145.1 slot for non-channelized SSB and CW use, with FM simplex relegated to outside the new one-megahertz band; 19 of the Northern California systems represented volunteered to move into the new band when it becomes available November 4.

FCC'S DECISION ON BANNING LINEARS for 10 meters and the Type Acceptance of Amateur of Amateur equipment has been put off until at least this month. The Commissioners have strongly supported Type Acceptance, now limited to only Amateur linear amplifiers, and that docket — 21117 — alone would have passed without difficulty. However, a decision has been made to permit oral arguments on both dockets.

POINT-OF-SALE CONTROL of Amateur transmitting equipment was stressed by both the ARRL and Drake at the September 15 Congressional hearings on rewriting the Communications Act. Dick Horner, E.F. Johnson's President, made a strong pitch for 220-MHz CB and drew some fire from the Amateur Radio representatives. Some observers felt the hearings were a bit disjointed, with our side making some good points but major emphasis on CB — particularly CB problems — rather than Amateur Radio.

A "TVI-PROOF" TV RECEIVER, built for the FCC by Texas Instruments, is reported to look very promising in preliminary tests. If the techniques TI has developed to reduce or eliminate TVI problems could be quickly adopted by the industry, a lot of the pressure for repressive rule making could be taken off the FCC. Amateur as well as CB and TV manufacturers are expected to be watching the Commission's tests with a great deal of interest.

DISTINCTIVE PREFIXES SUCH AS KG6 and KV4 are being discontinued for various Pacific and Caribbean islands. Instead of their present unique prefixes, all Pacific area U.S. Amateurs will be issued KH6 calls while those in the Caribbean will receive KP4 prefixes. Present holders of calls with the discontinued prefixes will, however, be permitted to retain them indefinitely — the change applies only to new applicants from those areas.

The Prefixes Involved include KG6, KS6, KB6, KJ6, KM6, KP6, KW6, KV4, and KC4 (Navassa). The reasons for the change include freeing up a large number of Amateur callsigns for future Amateur growth and reduction of the processing burden.

KANSAS CITY'S DISASTROUS FLOOD found Amateur Radio providing communications supporting police, fire, and other area relief agencies. The six area repeaters operated 'round the clock after 12 inches of rain flooded much of the city and drowned at least 23 people. WBØOAY, the emergency station located in the underground disaster center in Lee's Summit, operated from the center's emergency power system while providing key liaison between the Amateur and municipal communications channels. Area CBers were also valuable contributors to the volunteer communications effort, providing local communications which were relayed to the civil authorities via Amateur repeaters.

PRE-1917 AMATEUR LICENSEES must apply for "Grandfather" credit toward an Extra Class license before next March 1, after which it will no longer be offered. Grandfather credit has been available for quite a few years but no one has claimed it for some time.

General Class Amateurs licensed before 1933 would be grandfathered to Advanced if a petition filed with the FCC by Paul Halmbacher of Milwaukee were granted. Comments on the petition, RM-2884, may be filed through October 19.

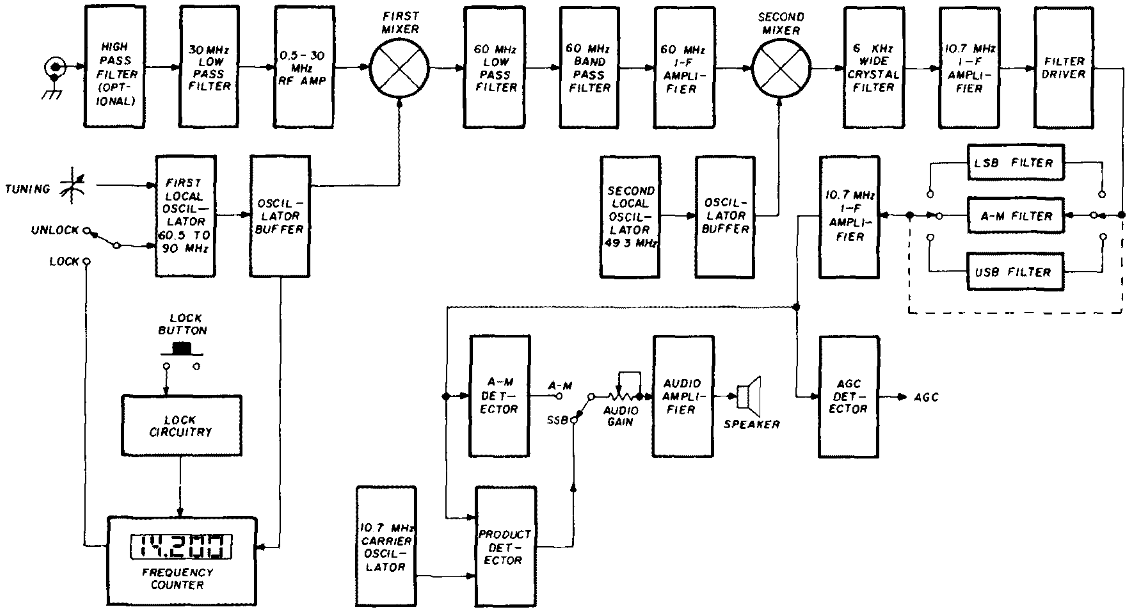



**fig. 1. Block diagram of a high-performance communications receiver that tunes from 500 kHz to 30 MHz. The design features digital frequency readout, upconversion to 60 MHz, i-f selectivity at 10.7 MHz, and a unique digital control system to maintain frequency stability.**

**Choice of i-f.** There are a number of factors which influence the selection of the receiver's i-f, including image response, oscillator tuning range, and i-f rejection. If the chosen i-f is too close to the highest input frequency, signals will leak through the rf stages to the i-f amplifiers. An i-f that is too close to the desired rf input frequency also enhances undesired image response. Finally, as the frequency of the i-f is increased, the local oscillator frequency must also be increased — and it's more difficult to build a stable oscillator at higher frequencies.

If you wish to tune the complete band from 500 kHz to 30 MHz, the i-f must be placed outside the tuning range, preferably above 30 MHz (placing it below 500 kHz leads to undesired image response, as discussed previously). When choosing an i-f above 30 MHz, select a frequency where there are no high level rf signals (between two vhf television channels, for example). Furthermore, choose an i-f which has no high level rf signals at the image frequencies. A receiver with an i-f at 40 MHz which tunes from 500 kHz to 30 MHz, for example, requires a local oscillator which tunes from roughly 41 to 70 MHz so images fall in the range between 71 MHz and 100 MHz; since this frequency range covers television channels 5 and 6 as well as a good part of the fm broadcast band, this i-f is obviously not a good choice.

For a high-frequency general coverage receiver, an i-f at 60 MHz represents a reasonable compromise. It falls between television channels 2 and 3, and its image frequencies are between 120 and 150 MHz, bands occupied primarily by low-power aircraft communications and other relatively low-power radio services. This is the i-f I chose for the receiver described in this article.

A low-frequency oscillator, say 5 to 6 MHz, could be mixed with thirty different crystals to provide the 60 to 90 MHz injection signal, but this would cause several problems: the 5-6 MHz signal falls within the passband of the receiver and many crystals are required. In addition, a large number of filters would be required to eliminate unwanted mixer products.

Another approach would be to build a 60 to 90 MHz oscillator which is electrically and mechanically stable, and provide an electronic lock circuit to keep the oscillator on frequency. This can be done with a

**table 1. Performance specifications of the general-coverage communications receiver designed by W6URH.**

| | |
|---|---|
| Frequency range | 500 kHz to 30 MHz continuous change |
| Sensitivity | 1 μV for 10 dB signal-to-noise ratio (6 kHz bandwidth) |
| Image rejection | −87 dB |
| I-f rejection | −85 dB |
| Out-of-band IMD | −35 dB |
| Cross modulation | −25 dB |
| Agc range | 100 dB for less than 10 dB change in audio |
| Audio distortion | less than 3 per cent |
| Audio output | 3 watts |
| Power consumption | approximately 50 watts |

up/down frequency counter if you add digital storage — a 7475 quad latch for each stage of the counter. When the lock button is pushed, the digital storage or memory is frozen. The counter continues to count the local oscillator and compares the measured frequency with that stored in memory. A dc voltage which corresponds to the error tunes a varactor diode in the oscillator, thus keeping the local oscillator locked on the frequency contained in memory. This is essentially the system used in this receiver. A block diagram of the receiver is shown in **fig. 1**; operating specifications are listed in **table 1**.

**Front end.** There are two filters preceeding the rf amplifier: a 2-MHz highpass filter, and a 30-MHz lowpass filter. The 2-MHz highpass filter minimizes cross modulation of signals above 2 MHz caused by strong signals in the a-m broadcast band; it is automatically disconnected below 2 MHz by logic circuitry in the counter. Response of this filter is down 0.5 dB at 1.9 MHz, and 60 dB down at 1 MHz. If the receiver is not located within several miles of one or more 50 kW broadcast transmitters, this filter may not be required.

The 30-MHz lowpass filter attenuates signals above 30 MHz, particularly signals at the image frequencies between 120 and 150 MHz. The response of this filter is 45 dB down at 50 MHz, 70 dB down at 100 MHz, and at least 60 dB down out to 460 MHz, the upper frequency limit of the test equipment I had available.

The rf amplifier provides about 10 dB gain and will handle signal levels as high as 1 volt rms without appreciable overload or cross modulation. A double-balanced mixer, MX101, mixes the incoming signal with the 60.5-90 MHz first local oscillator to provide the 60-MHz first i-f. Since the 60-MHz lowpass filter is not resistive, a 200-ohm resistor at the output assures termination at all frequencies; 60-MHz lowpass and highpass filters follow the mixer.

The mixers I used require +7 dBm (5 mW) of local oscillator drive; I used +13 dBm (20 mW) oscillator injection to increase the dynamic range of the receiver. If even greater dynamic range is required, high-level double balanced mixers such as the Mini-Circuits SRA-1H should be used: this mixer requires an injection level from +15 to +22 dBm (32 mW to 158 mW).

**First local oscillator.** I tried several oscillator circuits including the Colpitts and the popular grounded-base oscillator which is popular in TV receivers and fm tuners; all suffered from problems with power supply hum and frequency drift. The modified Vackar oscillator I used in the final design is very stable and has negligible power supply hum. During the past half hour, as I was working on this article, I have been listening to a ssb net on 20 meters — no retuning of the receiver was required.

In the first local oscillator (see **fig. 2**) Q201 is a free running fet oscillator which is tuned by C210 from 60.5 to 90 MHz. The oscillator is followed by a source follower, Q202, and a grounded base amplifier, Q203. The ferrite bead in the base of Q203 prevents it from oscillating at about 600 MHz; this bead should not be omitted here or in other places where a 2N5179 is used as an amplifier.

Q301 provides gain and isolation between the oscillator and the first mixer; Q302 and Q303 provide isolation *from* the counter. Isolation is required because signals generated within the counter tend to leak backwards into the first mixer and produce spurious signals.

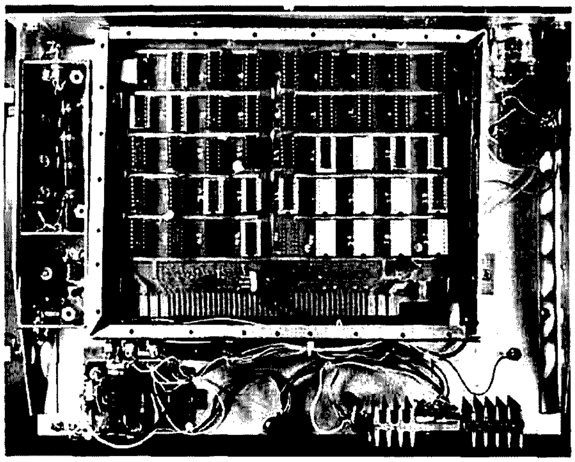
**60-MHz amplifier.** C401, C402, and L401 transform the 50-ohm output of the 60-MHz bandpass filter to about 1500 ohms. Q401 provides about 10 dB of gain to compensate for loss in the first mixer and the two 60-MHz filters; the gain is set by R401. MX401 combines the 60-MHz i-f with the second local oscillator signal at 49.3 MHz to produce the 10.7-MHz second i-f. Q402 provides isolation and gain from the second local oscillator to MX401. The output of MX401 is terminated with 200 ohms, like the first mixer, and transformed up to 1600 ohms by C403, C404, and L403 to drive a 6-kHz wide crystal filter. The output of the crystal filter drives Q403 which provides gain.

In the second local oscillator Q501 is a grounded-base crystal oscillator operating at 49.3 MHz; Q502 provides gain and isolation. Bandpass filter FL501 attenuates *harmonics from* the oscillator.

**Information filters.** Q601 provides matching to the filters. The filters are switched on by applying +18 volts through S1A to the diodes associated with each filter. The drive and termination values should be those specified by the filter manufacturer.

The USB and LSB filters I used required 500 ohms, and the a-m filter required 1600 ohms. To provide uniform gain the output of the a-m filter is attenuated with an 1100-ohm resistor.

**10.7 MHz i-f.** U701 provides further i-f gain; Q701 provides 10 volts p-p voltage to drive the detectors; CR701 and CR702 comprise the agc detector. The output of the agc detector is set at approximately +3 volts and goes negative when a signal is applied. This detector has fast attack, about 10 ms, and slow decay to eliminate the necessity of shutting off the



Top view of the general-coverage receiver, showing layout of the logic ICs for the digital readout and digital frequency control. The second local oscillator and filter FL501 are in the shielded compartment to the left. The S-meter is in the upper right-hand corner.

agc and using an rf gain control for ssb and CW reception; CR703 and CR704 are the a-m detector. Q801 is a 10.7-MHz crystal-controlled oscillator which provides carrier for Q802, the product detector. Oscillator drive is about ten times greater than the signal level.

In the audio amplifier stage R1 is the audio level control which is connected to one half of an LM379 audio amplifier; the amplifier provides about 3 watts of audio. In the metering circuit S901 switches between the S-meter and vco error voltage. Q901, a source follower, provides a current source for the S-meter.

## frequency counter

One half of an MC1004, U8, is a 1-MHz oscillator; the other half is used as a buffer. Transistor Q10 translates the ECL level from the MC1004 to TTL level for the 7490 divider chain. Note that only this ECL chip is operated between +5V and ground; all other ICs use −5V and ground. U18, U27, U36, and U45 divide the 1-MHz signal down to 100 Hz; U44 divides 100 Hz by 12; this counter is reset to four rather than zero because it must always go to count 15 in order to reset. Counts 4 through 13 provide 10 Hz or a 0.1 second counting period. Count 14 is decoded by U26, a 7420, and used to strobe the latches. Count 15, from pin 15 on U44, provides reset or preset for the counters.

The counters, U10 through U15 and U34, have two modes of operation. In the *unlock* condition they up-count the frequency of the first local oscillator; in the *lock* mode the counters are preset to the number located in latches U1 through U6 and U16, and count down to zero. U34, an MC10137 up-down counter, has four operating modes which are determined by the voltages on pin 7 (S1) and pin 9 (S2)

| S1 | S2 | Operation |
|---|---|---|
| low | low | preset |
| low | high | count up |
| high | low | count down |
| high | high | hold (stop count) |

For the 74192 up-down counters to count up, the input signal must be connected to pin 5; for count down the input signal is connected to pin 4. Clear is determined by the level on pin 14; load or preset is determined by the voltage on pin 11.

**Unlock operation.** Signals from the first local oscillator are fed to U34. The U34 outputs (Q1 through Q4) are translated to TTL level by U25 and fed to U16. During the strobe period the display is updated; during the reset period U34 is reset. Gates U33 and U20 direct U34 to count up, hold or reset. The 1000- and 1200-ohm resistors convert the TTL output to ECL level. Since one input to each gate in U7 is zero during unlock operation, the counter will always reset to zero.

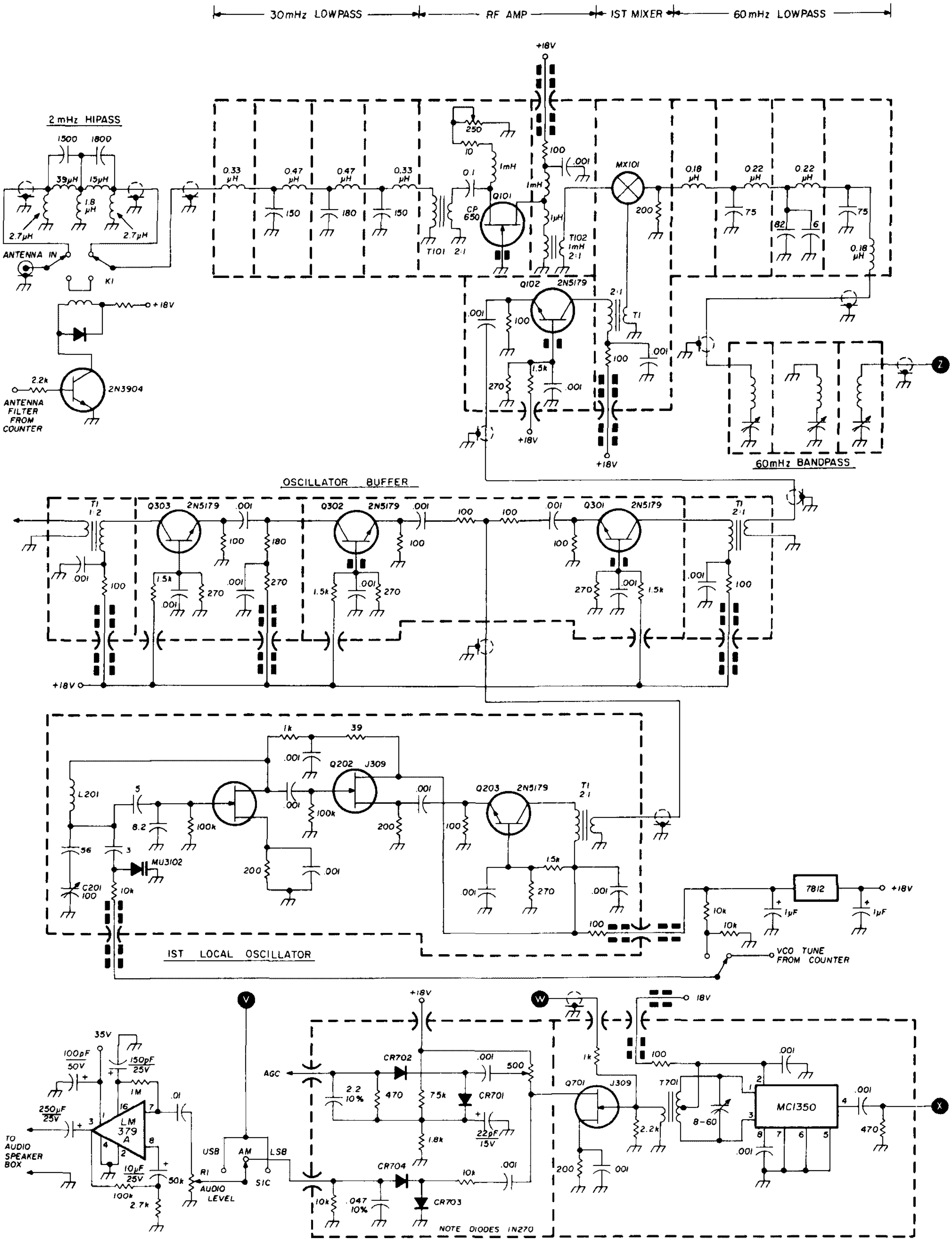
The output from U34 goes through a ECL to TTL level converter into U32, a one-shot multivibrator. The purpose of U32 is to widen the pulse being counted to about 50 ns. During unlock U23, pin 9 is high, and the signal being counted goes to U15, pin 5, count-up input. At the end of the count, during the strobe period, information stored in U6 updates the 100-Hz display. Counters U10 through U14 operate in a similar manner.

**Count period.** 10 units or 0.1 second. The counters count the input frequency.

**Strobe period.** 1 unit or 0.01 second. The counters hold their count, and this latest count is transferred into the display through the 7475 latches.

**Reset period.** 1 unit or 0.01 second. The counters are reset to zero and held there until the start of the next count period. Note that the 10-MHz display is connected to its latch, U1, differently than the others. This is done so the 10-MHz display will read the frequency to which the receiver is tuned rather than the actual frequency of the first local oscillator.

**Lock operation.** When the lock button is pushed, the circuit remembers the frequency of the first local oscillator in the 7475 latches, and when it drifts, brings it back to this frequency. Pushing the lock button sets an RS flip-flop which debounces the


30mHz LOWPASS
RF AMP
1ST MIXER
60mHz LOWPASS
2mHz HIPASS
ANTENNA IN
K1
ANTENNA FILTER FROM COUNTER
2N3904
T101 2:1
Q101
CP 650
T102 1mH 2:1
MX101
Q102 2N5179
60mHz BANDPASS
OSCILLATOR BUFFER
Q303 2N5179
Q302 2N5179
Q301 2N5179
Q202 J309
Q203 2N5179
L201
C201 100
MU3102
1ST LOCAL OSCILLATOR
7812
VCO TUNE FROM COUNTER
TO AUDIO SPEAKER BOX
LM 379 A
R1 AUDIO LEVEL
USB
AM
LSB
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AGC
CR702
CR701
CR704
CR703
NOTE DIODES 1N270
Q701 J309
T701
MC1350
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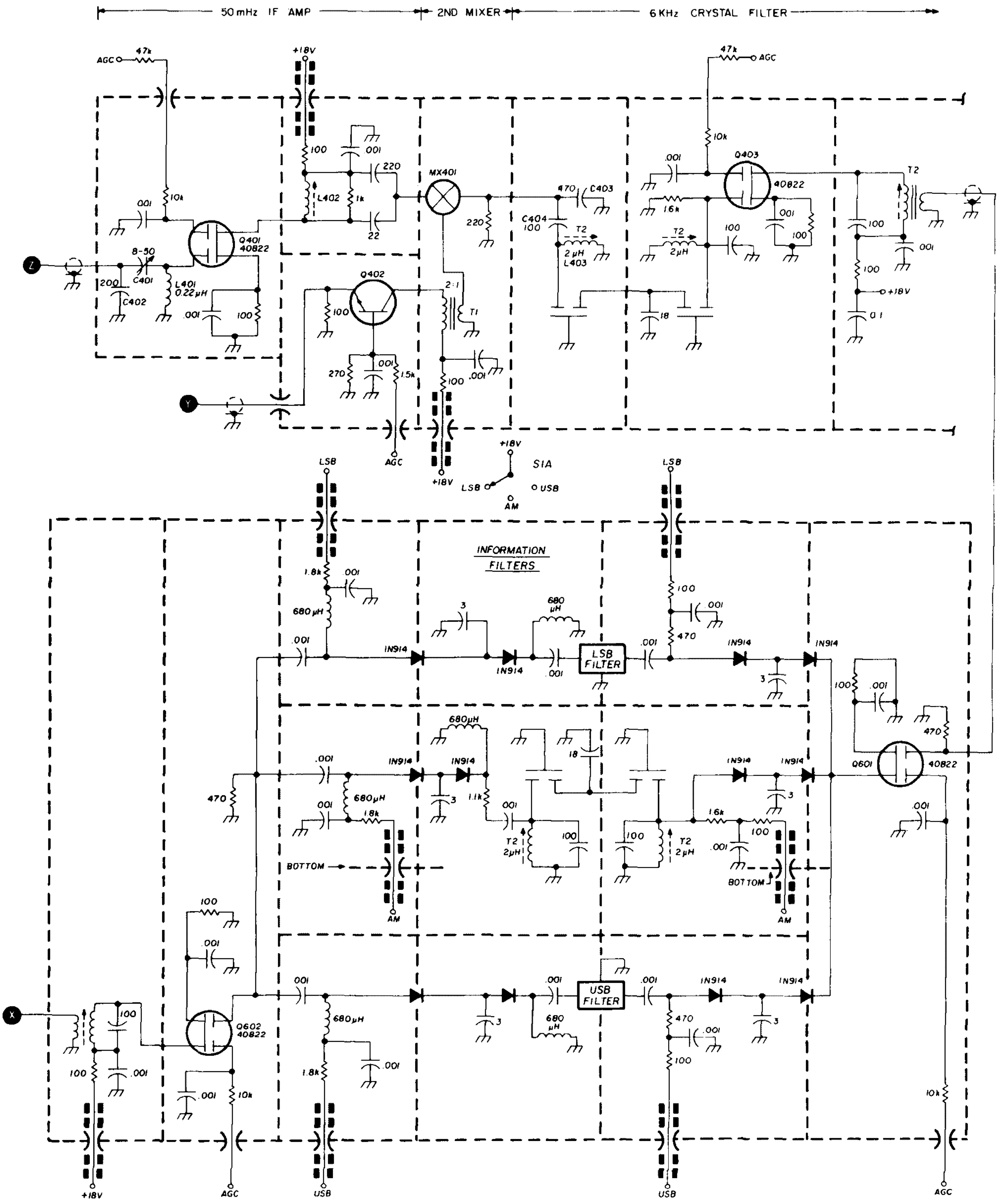
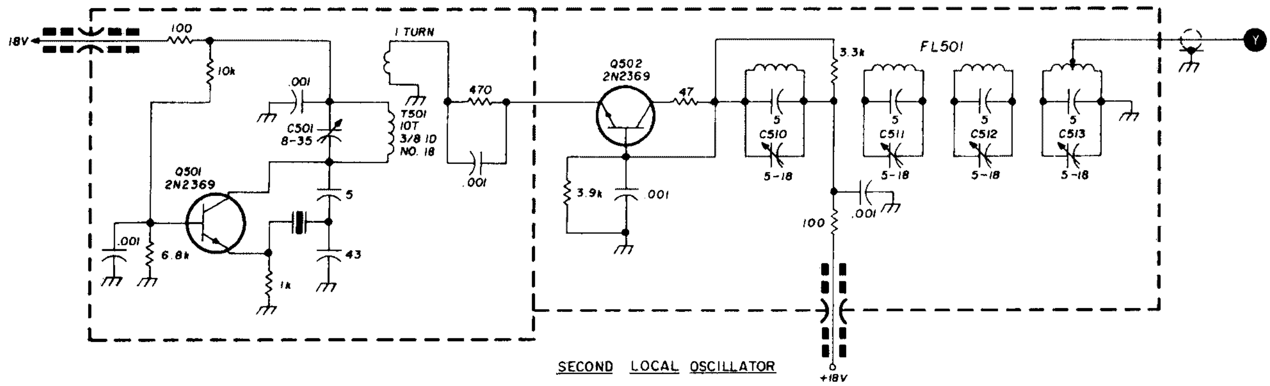
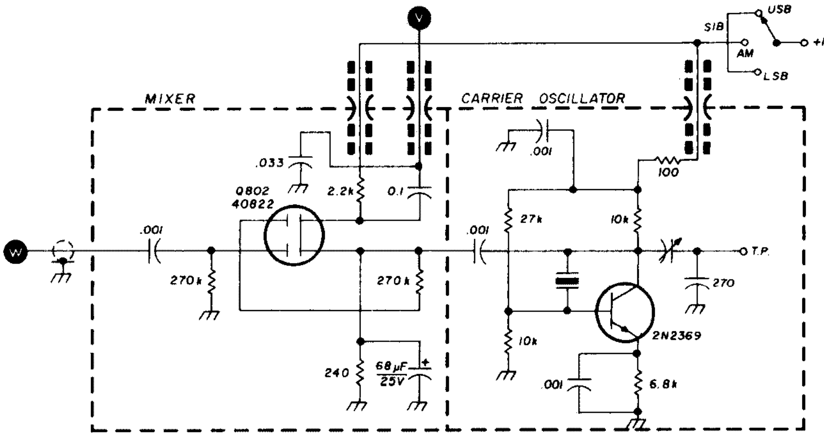


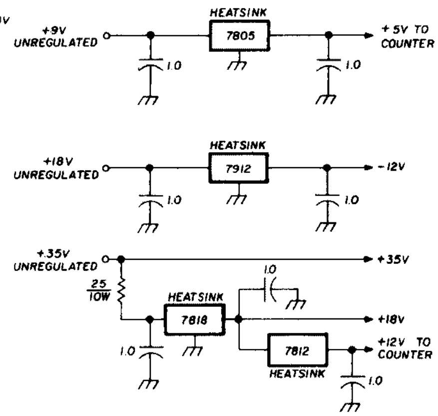

fig. 2. Schematic diagram of the high-performance general-coverage receiver. All feedthrough capacitors are 1000 pF; other capacitors between 1 pF and 2000 pF are silver mica. The 0.001 bypass capacitors are ceramic discs; capacitors marked 0.01 are ceramic discs; capacitors marked 0.01 to 2.2 μF are ceramic, monolithic, or tantalum types; larger value capacitors are electrolytics. All resistors are ¼ watt, 5 per cent, unless otherwise indicated.

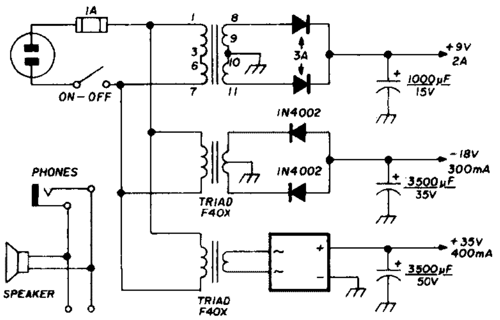



SECOND LOCAL OSCILLATOR

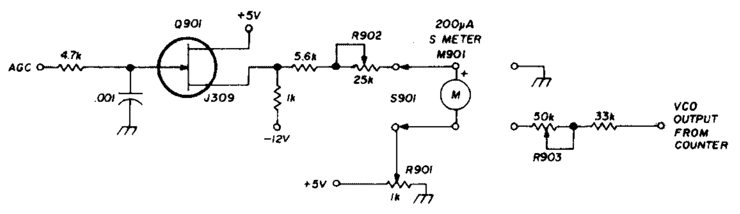



PRODUCT DETECT



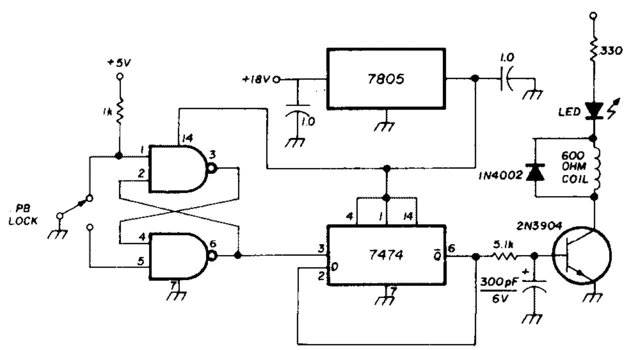




SPEAKER BOX




METER CIRCUIT

| | |
|---|---|
| FL501 | 4 windings no. 24 (0.5mm) wire on 1/2" (13mm) form, 1/2" (13mm) between windings |
| L402 | 7 turns no. 26 (0.4mm) on a 1/4" (6.5mm) slug-tuned form |
| MX101, MX401 | Mini-circuits SRA-1 double-balanced mixer |
| T1 | 4 turns no. 30 (0.25mm) trifilar wound on a ferrite bead |
| T2 | 2 µH. Turns ratio of 3.3:1 (transformers used by the author were unmarked and probably were intended for an fm i-f strip) |
| T101, T102 | Watkins-Johnson BT8. Construction as for T1 would probably be satisfactory |
| T501 | 10 turns no. 18 (1.0mm), 3/8" (9.5mm) diameter; secondary is 1 turn |
| T701 | Primary is 40 turns, secondary is 15 turns, no. 32 (0.2mm) wire on Micrometals T30-6 toroid form |
| Beads | Stackpole 57-180 or Amidon FB 43-101 |




LOCK CONTROL

pushbutton. Once the button is pushed and the RS flip-flop is set, it stays set even if the switch opens for an instant. The 7474 is connected as a divide by 2; each time the 7474 receives a positive edge trigger it changes state.

Signal is fed from the 7474 to the counter lock circuit. At the same time a relay is activated which switches the first local oscillator from a fixed voltage (developed by the two 10k resistors) to vco control. The 5.1k and 300 $\mu$F capacitor on the base of the 2N3904 transistor driving the relay ensure the oscillator is not put into lock until the proper frequency is stored in the 7475 latches.

Back in the counter U22 receives the lock signal and waits until the strobe period to send a lock command to the counter. The counters cannot be put into lock when they are counting because they would record a partial count; this would cause the first local oscillator to attempt to lock to a frequency other than the frequency to which the receiver was tuned. The latches now contain the frequency at the time the lock button was pushed. They will retain this number until the lock button is again pushed to unlock the counter.

The latches are held closed by changing the level on U33 pin 2 to a zero; U7 is activated. During the reset period U34, instead of being reset to zero, is preset to the number stored in U16 and counts down from the number stored in U16 rather than counting up. The 74192 counters are loaded with the numbers contained in their 7475 latches and count from that number to zero.

## error detector

U23, U19, and Q1 through Q5, a sample and hold circuit, provide compensation for oscillator drift. U23, an RS flip-flop, sets when it receives a borrow pulse from U10. The borrow pulse from U10 occurs when all counters have reached zero.

Operation when the oscillator stays on frequency, drifts up, and drifts down, is described below, but the examples are exaggerated. The first local oscillator, after a few minutes warm up, is sufficiently stable in unlock for at least 5 minutes of ssb reception without retuning.

**Oscillator frequency remains constant.** If the oscillator remained exactly on frequency, a borrow pulse would occur exactly at the end of the count period; U19A would not change state because U23 holds it in reset. U19B would not change state because it receives an $\bar{S}$ signal holding it in reset at the time it receives the signal from U23.

**Oscillator increases frequency.** If the oscillator frequency changed from 70 to 80 MHz, for example, the increased number of pulses would cause the counters to reach zero before the end of the count period; the borrow pulse would set U23. This would change the state of U19B and turn off Q2 which is normally held on by the $\bar{Q}$ output of U19B. This results in a positive-going signal at the collector of Q2 which turns on Q4. Q4 then partially discharges C1. This decrease in voltage passes through Q5, a source follower, causing an increase in varactor capacitance in the first local oscillator, thus reducing its frequency. At the end of count period U19B is reset by the S signal; U23 is reset during the reset period.

**Oscillator decreases frequency.** It should be noted that when the counter is in lock, the counters are allowed to continue counting during the strobe period. If the oscillator frequency was to go from 70 MHz to 60 MHz, U26 would not be set until sometime after the start of the strobe period. At the start of the strobe period U19B is held in reset. The U19A clock input receives the strobe pulse, thus changing its state and turning off Q1. This results in a positive-going signal which turns on Q3, increasing the charge across C1. This increase passes through Q5, the source follower, and on to the varactor diode in the oscillator. As the frequency of the oscillator increases the borrow pulse sets U23 and resets U19A; U23 is reset during the reset period.

## low-frequency detector

U28, a 7430 8-input NAND gate, looks at U1 and U2 — if the first local oscillator is between 60 and 62 MHz the output of U28 goes low. This turns off the transistor located at the 2 MHz highpass filter, which in turn deactivates relay K1 and disconnects the 2-MHz highpass filter.

## some initial problems

Initially the receiver was built on sheets of copper-clad circuit board with normal bypassing and decoupling, but without enclosing individual stages. The two mixers were on separate boards but completely exposed. When the receiver was together and working, there were two serious problems. First, without an antenna connected, at least 50 carriers could be heard in the receiver when tuning from 0.5 to 30 MHz. I first thought that this was caused by harmonics of the two local oscillators beating against one another and installed a filter at the output of the second local oscillator; this didn't cure the problem.

A 1000-MHz spectrum analyzer was then coupled into various points in the receiver. In the vicinity of the mixers, harmonics of the oscillator increased dramatically. This is due to the fact that mixers are switches being turned on and off by the local oscillator; switching generates square waves and square waves contain many odd-order harmonics. The rf and i-f assemblies were boxed up; amplifiers

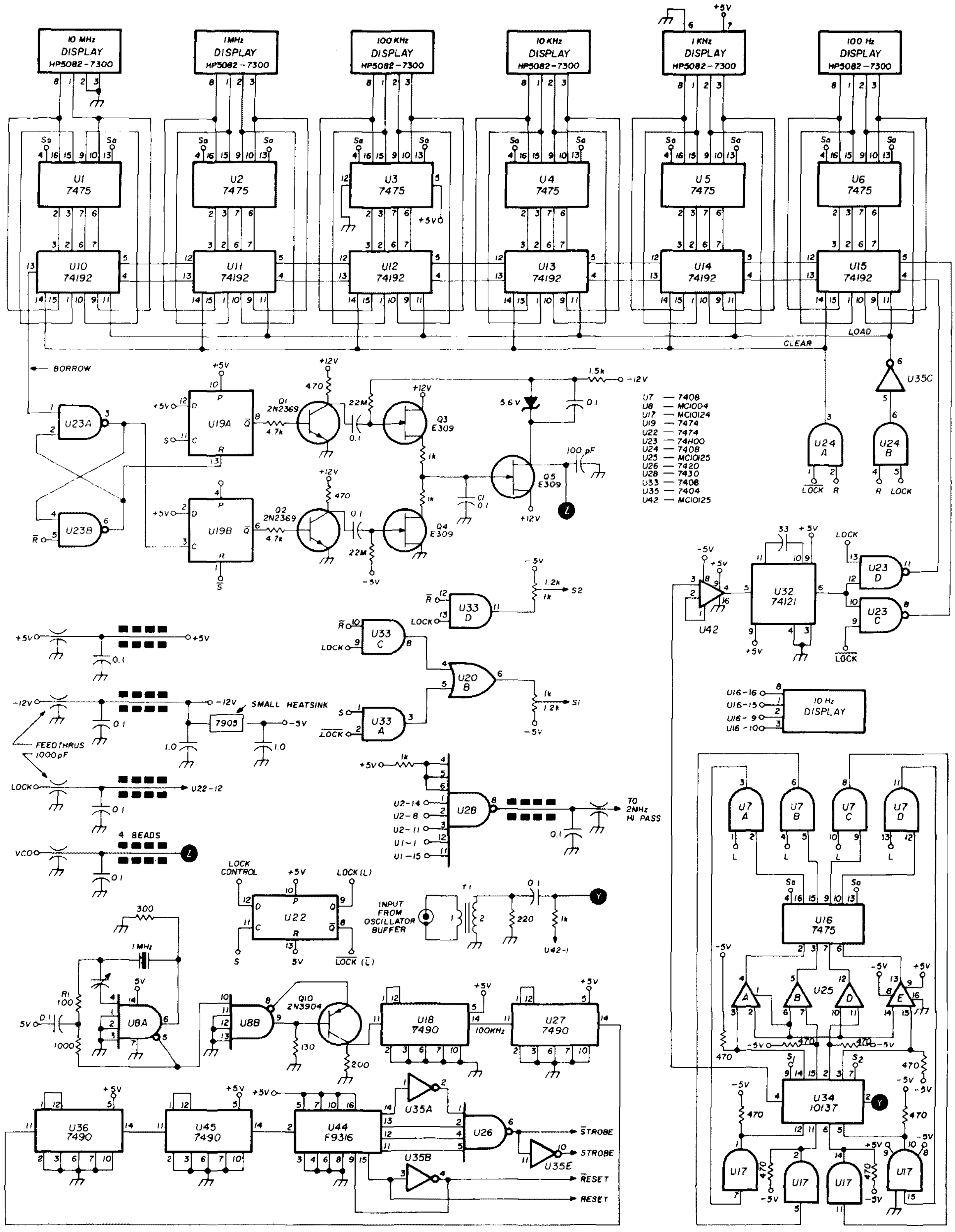



**fig. 3. Schematic for the digital frequency readout and digital frequency control system. Although not shown here, install a 0.1 μF bypass capacitor at every fifth IC ( + Vcc to ground). 22 μF, 15 volt bypass capacitors are installed on the +12, −12, +5 and −5 volt supplies. Digital displays are Hewlett-Packard type 5082-7300.**

with low reverse gain were installed close to each mixer local oscillator input. Finally, the circuitry was rearranged to reduce radiation from the rf and i-f ports of the mixers. This involved adding a 60-MHz lowpass filter after the first mixer because of blowby from the 60-MHz bandpass filter.

The results were gratifying. The strongest birdie left had an equivalent signal level of about 1 $\mu$V. It is very possible this birdie is the result of feedthrough from the first mixer to the second mixer. A 60-MHz crystal filter would both increase selectivity and reduce feedthrough, but such filters are relatively expensive. If you build this receiver without the enclosures and shielding, signals will abound, with or without an antenna!

The second problem partially involved the first problem. One of the first signals heard when an antenna was attached turned out to be the audio from television channel 44 about 30 miles (50km) away. It was caused by harmonics from one of the mixers in combination with leakage through the 30-MHz lowpass filter. The response of the original 30-MHz elliptical lowpass filter started dropping off at 30 MHz but went back up at about 200 MHz. An M-derived filter solved the problem.

All amplifiers were evaluated using the 1000-MHz spectrum analyzer. The 2N5179 amplifiers were found to be oscillating at about 600 MHz. In every case a ferrite bead in the base lead stopped this oscillation.

## construction

Most of the assemblies for this receiver are built on 1/16-inch (1.5mm) thick, single-sided copper-clad printed-circuit board. The same material is used for shields between successive stages. All stages except the last i-f have paper-thin copper foil (available at hobby stores) wrapped over the surfaces where the covers are attached. This reduces radiation into and out of the circuits, and minimizes TVI caused by the first local oscillator. All dimensions shown in the layout drawings (**figs. 4** through **12**) are *inside* dimensions. The lock control, audio amplifier, 2-MHz highpass filter, and S-meter circuitry are all built on PC boards without shielding or feedthrough bypassing.

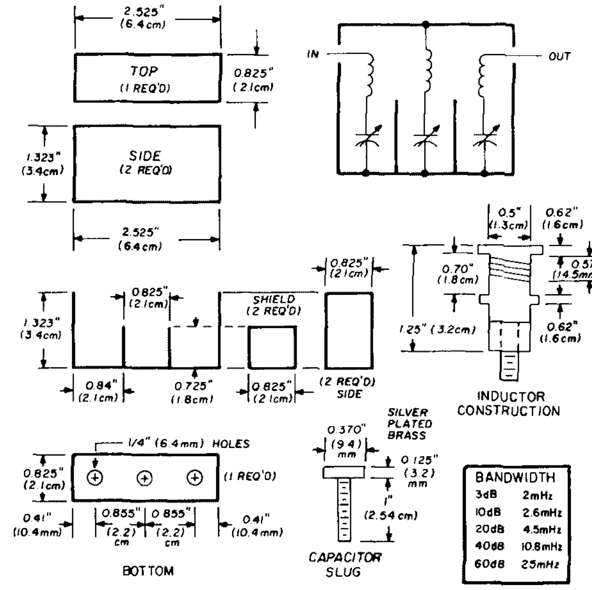



**fig. 5. Layout of the 60-MHz bandpass filter. The coil form is a ½" (13mm) slug-tuned National XR50. Tuning capacitance is provided by a slug mounted through the side of the shielded filter enclosure. Winding for each coil is 24 turns no. 30 AWG (0.25mm), spaced one wire diameter.**

**Counter.** The counter is built on a board with space for forty-five ICs. The analog circuitry is installed on a piece of copper-clad circuit board mounted underneath the counter and is surrounded by a 2-inch (50mm) high frame with perforated top and bottom covers. A wire mesh covers the display. All dc connections have four ferrite beads, and 0.1 and 0.001 $\mu$F feedthrough bypass capacitors. This is important because of the high level of rf radiation from the counter circuitry.

The heatsink measures 3-1/2 x 2 x 1 (9x5x2.5cm) and is mounted so it doesn't conduct heat to the main chassis. All high-power components — the counter and heatsink — are mounted on top of the chassis and air is allowed to convect around them for cooling. The circuitry below the chassis consumes relatively little power so heat generation is held to a minimum; this results in greater oscillator stability when the receiver is out of lock.

**Power supply.** The power transformers, rectifiers,

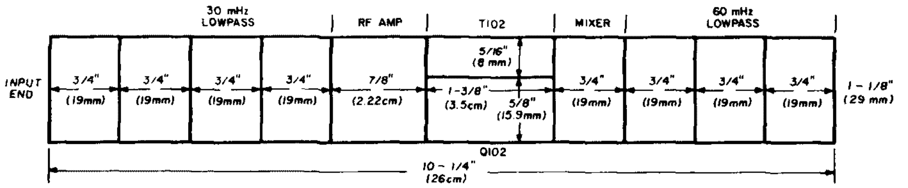



**fig. 4. Layout of the rf assembly (30-MHz lowpass filter through the 60-MHz lowpass filter, fig. 2). The unit is 1 inch (25mm) high. The area around the mixer is crowded in this layout; more space should be provided for this circuit.**

filter capacitors, and speaker are mounted in a separate box measuring 6 inches (15cm) on a side.

**Crystal filters.** The a-m filter is a Heath Dynamics A-5100. The ssb filters I used were manufactured by firms which do not sell filters in small quantities. Many of the 9-MHz crystal filters which are advertised for amateur use could be used with a few circuit changes.

**Other components.** The air variable capacitor used in the first local oscillator, and the coil forms used in the 60-MHz bandpass filter, will probably have to be obtained on the surplus market since these components have not been commercially available for several years. Many of the parts I used in this receiver came from my junkbox and those of other people so the original sources are unknown.

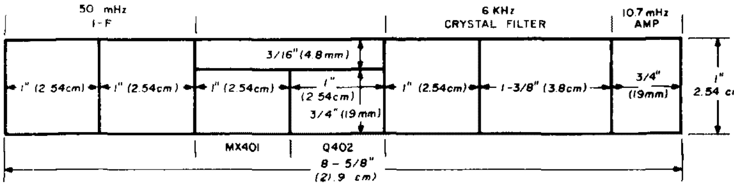



**fig. 6. Layout of the first i-f assembly (60-MHz i-f amplifier, second mixer, 6-kHz crystal filter, and 10.7 MHz i-f amplifier; fig. 2). The unit is 1-inch (25mm) high.**

**Chassis and front panel.** The chassis is 13-1/4 inches (33.7cm) wide, 10 inches (25.4cm) deep, and 2-3/4 inches (7cm) high. The front panel of the receiver is 13-7/8 inches (35.2cm) wide and 5-3/4 inches (14.6cm) high. These dimensions were chosen to fit into a cabinet I already had and are not critical.

## test and alignment

Following is a list of test equipment which I used to test and align the receiver. Use the test equipment you have available, but remember that the better quality equipment will provide better results.

1. Rf signal generator covering 2 to 60 MHz with a calibrated attenuator, 0.1 $\mu$V to 0.1 volt rms (Measurements 80 or equivalent).

2. Rf signal generator covering 500 kHz to 2 MHz, 0.1 $\mu$V to 0.1 volt rms (necessary only if you want to check sensitivity below 2 MHz).

3. Vacuum-tube voltmeter.

4. Rf voltmeter (such as the HP 410) or rf power meter (HP 430) to measure levels from zero dBm to +15 dBm.

5. Audio voltmeter with 30 mV full-scale sensitivity or oscilloscope with similar sensitivity.

6. Grid-dip meter.

7. Oscilloscope with minimum 10-MHz bandwidth (50 MHz bandwidth preferred).

**Power supply.** First check the input to the voltage regulator to verify that the voltages are approximately those shown on the schematic. Measure the output of all regulators for correct voltage output, and check to see that the supplies don't lose regulation at 105 Vac line voltage.

**Second local oscillator.** Use a grid-dip meter to verify that the 49.3-MHz crystal oscillator is oscillating *and* crystal controlled. Connect an rf voltmeter or power meter to the second mixer input and adjust bandpass filter FL501. The windings in this filter are critically coupled and require careful step-by-step tuning. Using a small screwdriver, short capacitor C512; adjust C511 and C513 for maximum deflection on the rf indicator. Short C511 and adjust C510 and C512 for maximum output. Now make slight adjustments to C510 through C514. This run-through may need to be done several times to optimize the filter. Replace the rf indicator with a counter and adjust C501 so the oscillator frequency is 49.30000 MHz. Adjust the J501 link so the level is about +13 dBm (about 1 volt rms). Connect drive to the second mixer. Install the cover and readjust C501 if necessary; replace the counter with the rf indicator and readjust C510 through C514.

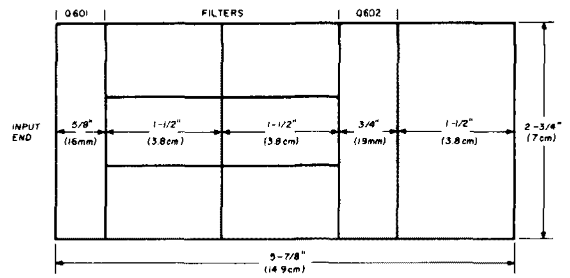



**fig. 7. Layout of the information filter assembly. The filters are mounted from the opposite side; standoffs are used to mount this unit to the chassis. The size and shape of this unit may be modified to accommodate filters which are available to the builder. Unit is 1-inch (25mm) high; last compartment is nearly empty.**

**First local oscillator.** Connect the counter to the first local oscillator. Squeeze or spread turns on L201 so the oscillator will tune from 60.5 to 90 MHz. The 56 pF capacitor may require a change in value to achieve this range (installing the cover will slightly detune the oscillator). Replace the counter with an rf

voltmeter or power meter and verify that the level is +13 dBm ±3 dBm from 60.5 through 90 MHz. Connect drive to the first mixer.

**I-f and rf alignment.** In **fig. 13** is a block diagram showing sensitivity at various points in the receiver. The alignment can be done one assembly at a time,

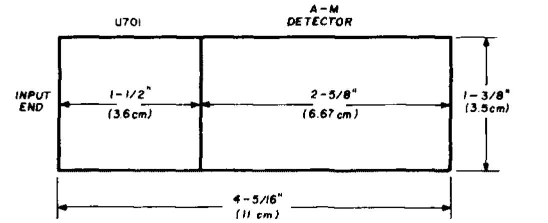



**fig. 8. Layout of the assembly containing the second i-f amplifier and product detector. Unit is 1 inch (25mm) high.**

but using this diagram and checking one stage at a time can provide some comparison of sensitivity.

Adjust the signal generator to 10.7 MHz, 30 per cent modulated with 400 Hz. Be sure the generator output impedance is 50 ohms; use a 6 dB pad if necessary. Connect an audio voltmeter (30 mV rms) to the output of the detector and turn the agc pot to minimum. Always maintain the generator output level so the voltage output at the detector is 30 mV or less.

The expected input to each stage, as shown in **fig. 13**, is provided by the rf generator; this level should provide 30 mV rms at the output of the a-m detector. The numbers shown are only approximate and will vary somewhat from receiver to receiver. In addition, the output of some rf-generators is not constant with frequency, so keep this in mind when making this test.

Connect the generator to the last i-f and terminate it with 50 ohms. Adjust the 60 pF variable capacitor for maximum indication on the output meter. Now move the generator to the input of information filter assembly and terminate in 50 ohms. Adjust all coils for maximum output. Go to USB and LSB and verify that the filters are working; slight adjustment of generator frequency will be required.

Switch back to a-m. Adjust the generator frequency to 60 MHz and connect it to the input of the i-f assembly. Adjust the coils and capacitor for maximum output. Now connect the generator to the input of the 60-MHz lowpass filter. Adjust the 60-MHz bandpass filter for maximum output; there is some interaction so it may require several adjustments. Connect the generator to the antenna input and verify that sensitivity is 1 μV or less for 10 dB signal-to-noise ratio.

**Product detector.** Connect the counter to the test point and adjust the trimmer for 10.700 000 MHz; verify LSB and USB operation.

**Agc and S-meter.** Connect the VTVM or dc-coupled scope to the agc line. Note that with no signal the agc line will be at about +3 volts. Now connect the signal generator to the receiver and tune both to about 10 MHz. Set the generator level at about 1 μV and turn the pot in the last i-f assembly so the agc voltage goes to about 2 Vdc. Turn the generator to 100,000 μV. If the receiver saturates, turn the agc level up until the receiver is out of saturation; the agc voltage should be about −0.25 volt. Disconnect the generator. Adjust R901 so the S-meter reads zero; connect the generator and adjust for 100 μV. Set the S-meter to S9 or wherever you would like it to indicate 100 μV.

## counter check-out

**Crystal oscillator.** Connect the counter to U18 pin 11 and set the trimmer so the oscillator is at 1.000 000 MHz. A fixed capacitor across the trimmer may be required to bring the oscillator on frequency. If the oscillator is not operating, increase R1 until oscillation starts. Turn the receiver ON and OFF several times to ensure that the oscillator always starts. If everything is working properly, the first local

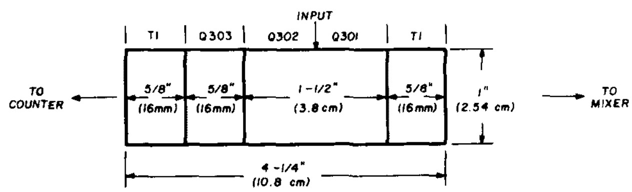



**fig. 9. Layout of the oscillator-buffer assembly. Unit is 1 inch (25mm) high.**

oscillator will now lock on frequency when the pushbutton is pressed. Turn S901 to vco error and adjust R903 for a midscale reading. Note that adjusting the tuning will cause the meter to move away from the center. At about 2 volts and 8 volts, the oscillator will go out of lock.

**Counter troubleshooting.** If the counter will not go into lock, the following test procedure may be used to localize the problem. Note that most problems are caused by wiring errors, faulty ICs, or bad sockets.

**table 2. ECL and TTL logic levels.**

| | logic zero | logic 1 |
|---|---|---|
| TTL | 0 to 0.8 V | 2 to 5 V |
| ECL | −1.5 to −5.0 V | −0.5 to −0.9 V |

All logic levels should be within the voltages shown in **table 2**. TTL for example should always be below 0.8 V or above 2.0 V; never in between.

Bring the counter out of lock; the meter will go to

almost full scale. Check for 100 Hz at U44 pin 2. Check for a 10 ms wide strobe pulse at U35 pin 11 and pin 12; check for similar reset pulses at U35 pins 3 and 4. Verify that U32 pin 9 changes state each time the lock button is pushed. See if U22 pin 12 is changing state; if not check the lock board. The 7400 RS flip-flop should change state each time the button is pushed and go back to its original state when the button is released. The 7474 should change state each time the button is pushed and stay until the button is pushed again.

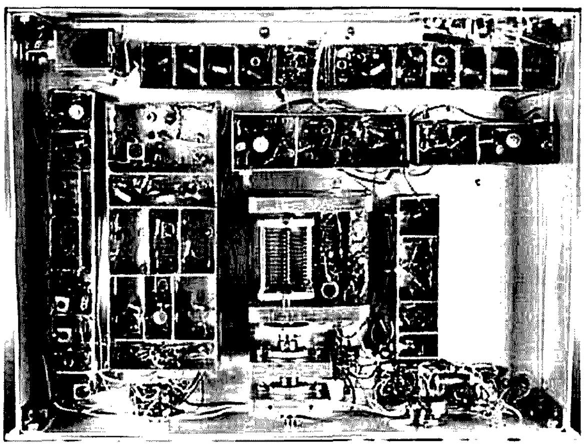
The frequency being counted comes out of U34 as a 20 ns wide pulse and is stretched out to 50 ns by U32. This 50 ns pulse becomes very difficult to see as it is divided down by the counters. Therefore, a low-frequency generator, at 1 MHz for example, makes investigation much easier.

Keep in mind that the most significant digit (MSD) has been modified to display receiver frequency so it will not correctly show the oscillator frequency.

Feed about 50 mV rms at 1 MHz into transformer T1 and look for 100 kHz pulses at U42 pin 3 (ECL level) and U42 pin 4 (TTL level) interrupted by strobe and reset pulses. The pulse from U32 pin 6 may be temporarily widened by adding a 200 pF capacitor between U32 pins 10 and 11. Check that the counter continues to count in lock.

Now remove U32 and connect the signal generator through the interface shown in **fig. 14** to U32 pin 6. Set the generator level to 3 or 4 V p-p; unlock the oscillator. Check for the pulse on pin 5 of U10 through U14. Push the lock button and check for the pulse on pin 4 of U10 through U14. By changing the frequency of oscillator, U23, an RS flip-flop, should be set momentarily and then reset with the reset pulse. U19 pins 6 and 8 should normally be high and go low depending upon the arrival time of the borrow pulse.

Connect a voltmeter to Q5 source and remove U19. Connect the signal generator through a TTL interface circuit to U19 pin 8. The voltmeter should indicate about +10 volts or more. Now connect the generator to U19 pin 6; the voltmeter should read +2 volts or less.



Bottom view of the communications receiver, showing the shielded compartments for the major assemblies. The enclosure at the top contains the rf amplifier assembly including the 30-MHz and 60-MHz lowpass filters. The unit in the top left-hand corner contains the 60-MHz bandpass filter. Along the left-hand side of the chassis is the 60-MHz i-f amplifier, mixer MX401, 6-kHz crystal filter, and 10.7-MHz i-f amplifier. The information filters are in the next enclosure to the right. The enclosure on the right-hand side of the vfo contains the buffer circuitry; the second i-f is center top, and the product detector is to its right.

**2-MHz filter driver.** Connect the oscilloscope to U28 pin 8. The output should be high when the receiver is tuned above 2 MHz, and low when the receiver is tuned below 2 MHz.

## receiver performance tests

There are several ways to check each parameter of receiver performance, and each of the tests described here can be accomplished in many other ways.

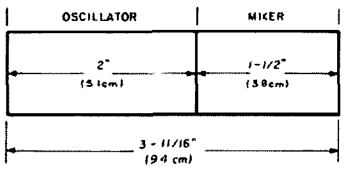



**fig. 10. Layout of the product detector which contains the carrier oscillator (Q801) and the mixer (Q802). The unit is 1 inch (25mm) high.**

**Signal-to-noise ratio** is a measure of how much signal is required to be 10 dB greater than the internally generated noise of the receiver. This test must be performed with the agc turned off (or set to minimum). Use caution not to overload the receiver. Tune the receiver and generator to 10 MHz (or other desired frequency) and set the modulation on the generator to 30 per cent. Connect an audio voltmeter to the a-m detector. Increase the signal generator output until the signal at the detector output is 10 dB above the receiver's noise floor. Output from the generator should be 1 $\mu$V or less; on my receiver it was 0.5 $\mu$V at 10 MHz, and 0.6 $\mu$V at 28 MHz.

**Receiver intermodulation.** There are two types of receiver intermodulation (IMD) tests. The first is to check the level of intermodulation products of signals outside the passband of the receiver; the second is to measure the intermodulation of two signals within the receiver's passband.

**Out-of-band intermodulation.** Tune the receiver

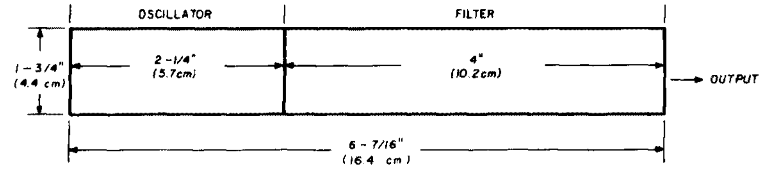



**fig. 11. Layout of the assembly containing the second local oscillator and the oscillator filter, FL501. This unit is 1¾ inches (45mm) high; the increased size is needed to minimize detuning of the filter. Holes, ¼ inch (6.5mm) in diameter, are provided in the top for tuning up the oscillator and filter.**

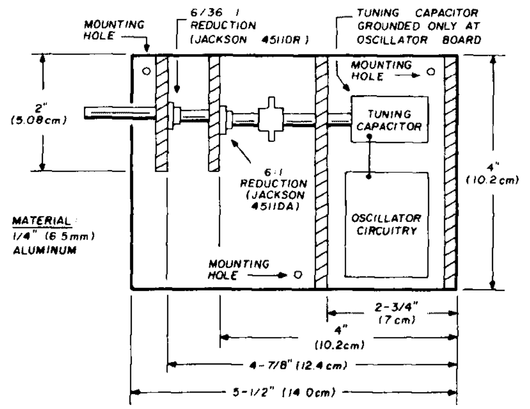



**fig. 12. Construction of the first local oscillator. The total height of this unit is 2¼ inches (67mm), including the ¼ inch (6.5mm) thick baseplate; a cover is provided to shield the oscillator circuitry and the tuning capacitor. This assembly is mounted on the main chassis with three 4-40 (M3) screws, and rubber grommets on the chassis. The tandem 6:1 and 36:1 Jackson Brothers drives provide an overall ratio of 216:1 at the front panel. The tuning capacitor is a 100 pF double-bearing air variable with brass plates. The oscillator coil is 6 turns no. 18 (1.0mm) enameled on a 3/8 inch (9.5mm) diameter form (slug removed).**

to 11 MHz and adjust the variable 100 dB attenuator pad so the signal level is at the noise floor of the receiver (**fig. 15**). Subtract 3 dB from this reading because two generators are used, and write down the result. Now tune the receiver to 10 MHz and reduce the attenuator level so an audio voltmeter at the detector indicates the same as the first reading. The difference should be 80 dB or more. In my receiver it was 85 dB.

**In-band intermodulation.** Tune both signal generators into the passband of the receiver, signals spaced 2 kHz (**fig. 15**). Connect a spectrum analyzer to the last i-f, or a wave analyzer to the a-m detector. Note the difference between one generator and the first product removed from this generator signal displayed on the spectrum analyzer, or the difference in level between the 1 and 2 kHz signals on the wave analyzer. The difference should be 20 dB or more for 10 per cent distortion, which is acceptable for radio communications. The measured difference in my receiver was about 35 dB.

**Agc range.** This is a measure of the change in audio output for a level change of the incoming rf signal. When changing the signal generator level from 1 μV to 100 mV the audio level in my receiver changed less than 10 dB (this measurement will vary with the setting of the agc control).

**I-f rejection.** This is sometimes called i-f blowby and indicates how far down leakage is at the first i-f with reference to the operating frequency. Tune the generator to 60 MHz and note the generator level for a 10 dB signal-to-noise ratio. Now tune the generator and receiver to 10 MHz and note the generator level for a 10 dB signal-to-noise ratio. The measured difference in my receiver was 85 dB.

**Image response.** This is a comparison between response to a desired in-band signal and its image. In the receiver described here a desired signal at 10 MHz has an image at 130 MHz. The measured difference in input level for the same output was 87 dB.

**Cross-modulation.** For this test two signal generators are connected as shown in **fig. 16**. One generator is tuned to 11 MHz and the output level is set to 200,000 μV, 30 per cent modulated; this is the interfering signal. The other signal generator is tuned to 10 MHz and the output level is set from 30 to

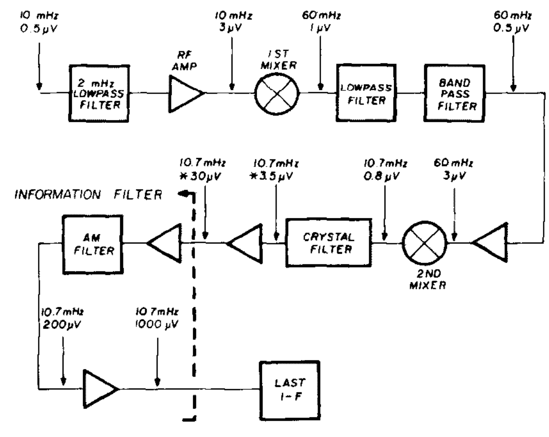



**fig. 13. Sensitivity at various points in the general-coverage communications receiver. Alignment can be done one *assembly* at a time, but using this diagram and checking one *stage* at a time provides a comparison of sensitivity. When making tests at the points marked with an asterisk, terminate the generator with a 50-ohm resistor. Turn the agc to minimum for all tests.**

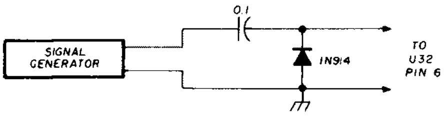



**fig. 14. Interface circuit for connecting the signal generator to U32 when troubleshooting the counter circuit.**

100,000 $\mu$V, 30 per cent modulated. The receiver is tuned to 10 MHz. In this receiver, modulation from the interfering signal is 25 dB below that of the desired signal when the level of the desired signal is between 30 and 100,000 $\mu$V.

## receiver performance

One of the best tests of receiver performance is to check to see at how many places on the dial a given

**table 3. List of high-power medium- and high-frequency transmitting stations located near W6URH.**

| station | distance miles | distance kilometers | power (kW) | frequency (MHz) |
|---|---|---|---|---|
| KGEI I | 6 | (10) | 250 kW | 6 - 15 MHz |
| KGEI II | 6 | (10) | 50 kW | 6 - 15 MHz |
| KNBR | 6 | (10) | 50 kW | 680 kHz |
| KLHI | 5 | ( 8) | 10 kW | 1550 kHz |
| KFS | 3 | ( 5) | 3 kW | 6348 kHz |
| KFS | 3 | ( 5) | 12 kW | 6365 kHz |
| KFS | 3 | ( 5) | 12 kW | 8444 kHz |
| KFS | 3 | ( 5) | 12 kW | 8558 kHz |
| KFS | 3 | ( 5) | 12 kW | 12.695 MHz |
| KFS | 3 | ( 5) | 12 kW | 12.844 MHz |
| KFS | 3 | ( 5) | 12 kW | 17.026 MHz |
| KFS | 3 | ( 5) | 12 kW | 17.184 MHz |
| KFS | 3 | ( 5) | 12 kW | 22.425 MHz |
| KFS | 3 | ( 5) | 3 kW | 22.515 MHz |
| KGO | 2.5 | ( 4) | 50 kW | 810 kHz |
| KDFC | 2 | ( 3) | 10 kW | 1220 kHz |

radio station can be heard — hopefully there will be only one. To make this test you have to be within ten miles of a broadcast or short-wave transmitter run-

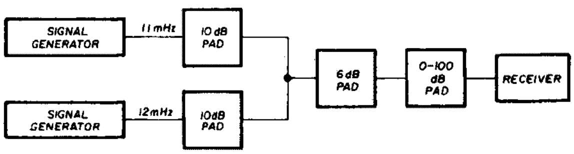



**fig. 15. Test set-up for checking intermodulation distortion. See text for test procedure.**

ning 1000 watts or more. I have sixteen such transmitters within 6 miles of my home operating between 680 kHz and 22 MHz (see table 3). Before I installed the 2-MHz highpass filter in the receiver, a-m broadcast stations could be heard at a low level as I tuned across the short-wave bands. KGEI I (250 kW) and KGEI II (50 kW) cannot be heard on frequencies other than their normal operating frequencies. Station KFS (up to 12 kW) has never been heard on frequencies other than those listed in table 2. I have heard harmonics of KDFC and KGEI but this is not a fault of the receiver.

I built this receiver for two reasons: to see if it was possible to build a practical receiver with up conversion — and to see if it was feasible to build a practical up-conversion receiver with frequency lock. Both have been proven to my satisfaction. However, there are several useful features which are not built in, including a noise blanker, receiver incremental tuning (possibly with a varactor), notch filter, audio filter for CW reception, and receiver muting (or send/receive switch). Any or all of these features could be added by the experienced builder.

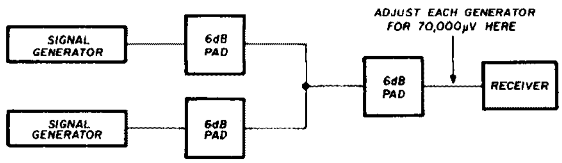



**fig. 16. Test set-up for measuring receiver cross-modulation. See text for test procedure.**

There are no plans for making printed-circuit boards for this receiver because standard PC construction would produce many unwanted spurious signals in the rf and i-f stages. PC construction might also lead to TVI from the first local oscillator.

I would like to thank Fred Scholtz, K6BXI, for help in the preparation of the article, and Marvin Kolber, K6PJU, for taking the photographs.

I would like to hear from anyone who makes any improvements in this receiver without degrading its overall performance.

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**ham radio**

# noise blanker design

A discussion of the design requirements for noise blankers which will effectively eliminate high amplitude, low repetition rate noise

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**The principles of noise blanking** are not new. The first description of the idea was published by Lamb[1] in 1936; however, there are subtle design considerations that have been overlooked in some previously published designs. This article will try to explain some of these considerations and the trade-offs that accompany a practical design. In addition, a brief description of a working circuit, used in one of my receivers, is given. First, however, you have to understand what noise blanking can and cannot do.

Under most conditions, noise blanking can minimize the effects of short duration, high amplitude, low repetition rate noise on a desired signal. Examples are some automobile ignition noise, certain electrical arcing noise due to power lines, and make or break switching.

Noise, with the opposite characteristics, long duration, low amplitude and high repetition rate, is difficult to control. This category includes lightning crashes, brush arcing, some powerline noise, and receiver generated thermal noise. These types of noise, as well as most others, are best dealt with at the source, if possible, or by minimizing them with other techniques such as lower noise amplifiers and directive antennas. (An article by Nelson[2] is an excellent source of information on noise sources.)

The reason a noise blanker is ineffective on lower amplitude and long duration pulses can best be understood by remembering that it operates by first sensing the presence of a noise pulse and then silencing the receiver for the duration of that pulse. The sensing operation requires discrimination between the signal and noise. Obviously, the greater the ratio between the two, the easier discrimination becomes. Silencing of the receiver is only permissible if the duration does not become so excessive that intelligibility suffers.

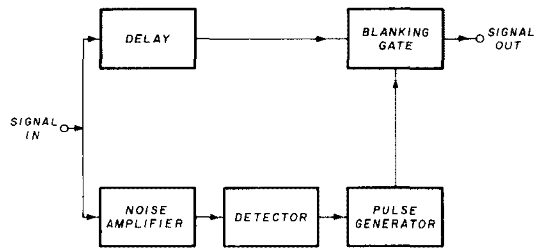


**fig. 1. Block diagram of a typical noise blanking circuit. The signal is split into two paths, the noise or control signal plus the original communications signal. The delay is introduced to ensure that the two signals arrive at the blanking gate simultaneously.**

## basic circuit

Now that the limitations are understood, let's examine a typical system. As shown in **fig. 1**, the incoming path is split into two channels. One channel is referred to as the main channel; the other, the noise channel. (The noise channel also contains signal, but the terminology is by convention.)

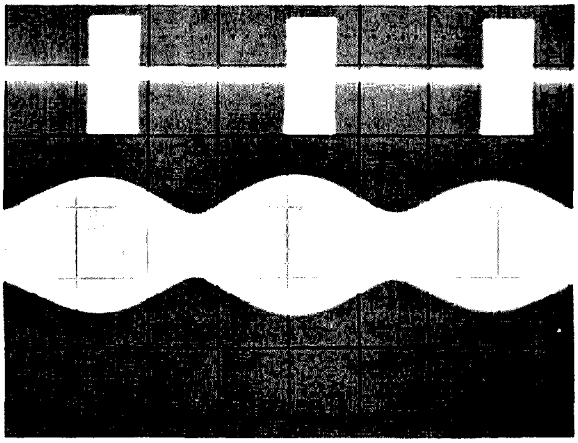

**fig. 2. Example of the effects of applying a noise pulse to a relatively narrow i-f filter. The upper trace shows an applied pulse while the bottom shows the output from the filter. Both traces were recorded at identical sweep speeds.**

The main channel is sometimes delayed, then passed through a gate and on to the rest of the receiver. In the noise channel, the noise and signal are amplified, and the noise impulses detected with the detector output are used to trigger the pulse generator. The pulse generator forms a signal of proper amplitude and polarity to cut off the gate for the duration of the noise impulse.

With the fundamentals behind us, look at some of the finer points. First of all, you must decide where, in the receiver, to place the blanker. For one thing, the blanking must be done prior to the narrow bandwidth i-f filter. The reason for this can be seen by examining **fig. 2**. The top trace of the photograph shows a simulated noise impulse which was applied to a mechanical filter with a 2-kHz bandwidth; the bottom trace is the output from the filter. While this is an extreme example of pulse stretching due to filter ringing, it shows the necessity of blanking at a point of wider bandwidth. However, it must be done before strong out-of-band signals become a problem. I have found a bandwidth of 50 to 100 kHz to be a reasonable compromise.

Another factor to be considered is amplifier overload. If you wait until the signal has passed through several amplifiers before blanking, the amplitude of the noise impulses may be high enough to have already overloaded one or more stages. The effects of nonlinear amplifiers are well known and need no further discussion.

So far, everything seems to indicate blanking very near the antenna. But, before going any further, take a look at the requirements for some of the other circuitry.

**Noise amplifier.** The input must be amplified to a level high enough to operate the threshold detector. Since the threshold point is generally in the range of

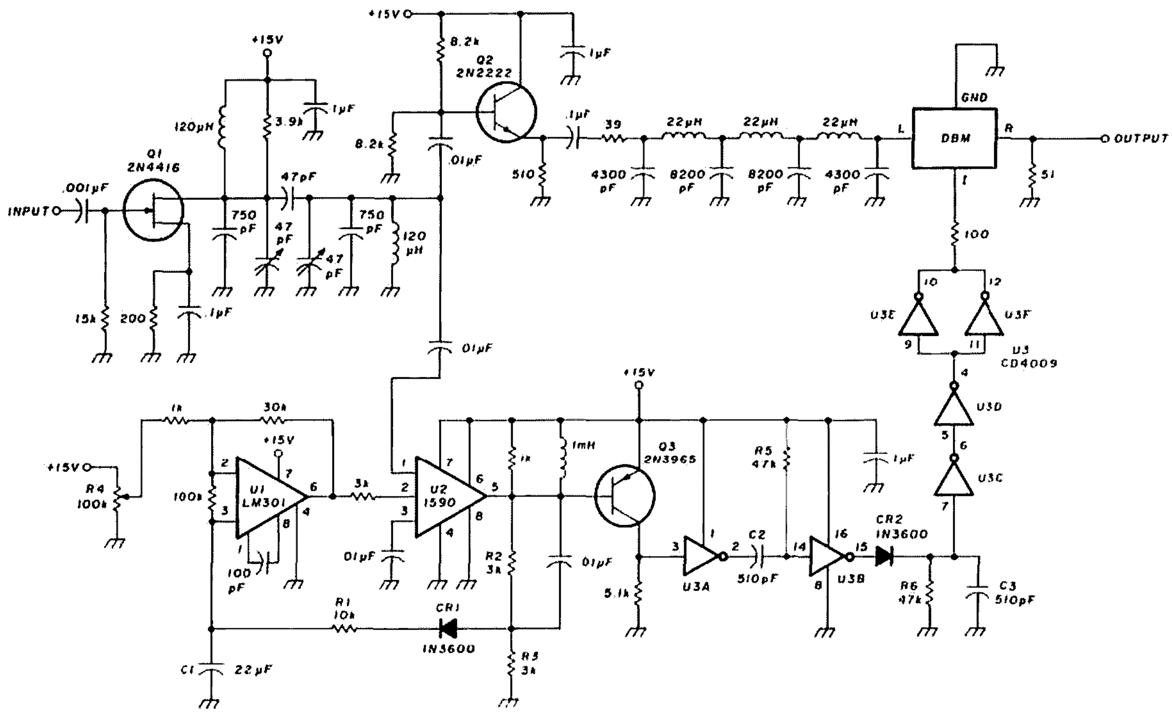


**fig. 3. The final design of the noise blanker as applied to the author's receiver.**

0.5 to 1 volt, the required gain may be extremely high if the input is small. This suggests placing the blanker at a point of high signal level. Again, there are two conflicting requirements and compromise is necessary. Also, for reasons of simplicity, the noise amplifier should be fixed-tuned which means it must be placed somewhere between the mixer and the narrow i-f filter. The bandwidth of the amplifier must, of course, be great enough to accurately follow pulse rise-time and minimize delay. If, as in my case, agc is used to obtain automatic threshold adjustment, the amplifier should maintain all its desirable characteristics with agc applied.

**Detector**. The ideal detector would have a very definite threshold below which it has no output and above which it has a large output. The response time must be very fast. There are many regenerative types of circuitry that would work, but I have found the simple transistor design used in the sample design (**fig. 3**) to be adequate.

**Pulse generator**. The function of the pulse generator was outlined earlier. Since the requirements will depend on the type of gate used, one circuit will not satisfy every need. A design that may come close is the retriggerable, one-shot multivibrator using CMOS ICs.[3] The retriggering action inhibits the gate for the duration of the noise pulse and then recovers very quickly. Risetime is relatively fast, but not so fast as to cause excessive transients and ringing. The voltage swing (0 to +15V) is high enough to operate most blanking gates. If required, CMOS gates may be paralleled for additional current capability.

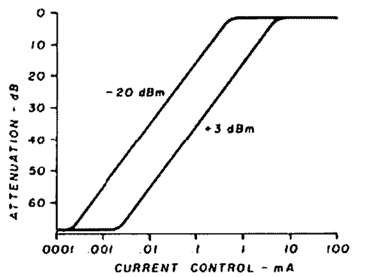


**fig. 4. A double-balanced mixer can be used as a current-controlled attenuator. This example shows the required current for two different applied power levels.**

**Delay network**. Since it takes a finite amount of time to amplify, detect, and form a pulse, a commensurate time delay should be introduced in the main channel to insure coincidence between the noise impulse and the blanking pulse. This delay is admittedly hard to come by at the higher frequencies and is left out in many designs. For lower frequencies the phase shift, through lowpass filters, is a reasonable method of introducing an apparent time delay. The amount of delay, $t_d$, can be calculated from:

$$t_d = \frac{\theta}{360 \cdot f}$$

where

$\theta$ = the phase shift in degrees
$f$ = the frequency of operation.

**Blanking gate**. Last, but perhaps most important, is the selection of a suitable gate. The characteristics of

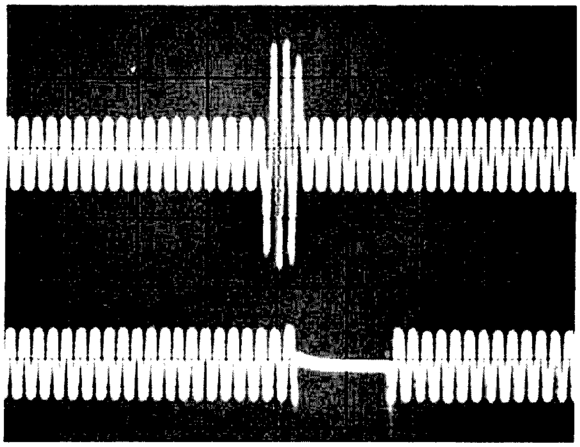

**fig. 5. The bottom trace of this photograph shows the output from the gate after a noise pulse has been blanked. The original pulse (top trace) had a 10 dB noise-to-signal ratio.**

an ideal gate are: zero insertion loss when on, infinite insertion loss when off, and no feedthrough of any switching transients to the output. This last point is extremely important. Some switching circuits, while doing a good job of cutting off the signal, can generate transients of larger amplitude than the original noise pulse!

I have tried many different types of gates, series and shunt diode bridges, switching fets, bipolar transistor switches, and even a double-balanced mixer (DBM) operated as a current-controlled attenuator. After evaluating all of the different possibilities, I settled on the DBM because of its good performance and simplicity. For those not familiar with this application of the DBM, **fig. 4** is a plot of attenuation vs control current at two different signal power levels.

## practical circuit details

**Fig. 3** is the complete circuit diagram of my blanker design. This particular circuit is installed in a modified Collins ARR-41 receiver. It is inserted

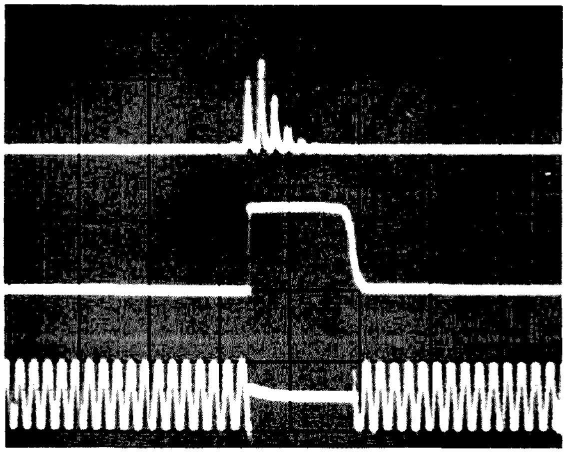

**fig. 6. Additional points in the blanker are shown in this photograph. The upper trace is the collector of Q3 while the middle trace is the output of the one-shot multivibrator. The bottom trace is the same input as in fig. 5.**

between the plate of the second mixer and the first i-f amplifier. The i-f is at 500 kHz and the signal level is a few millivolts.

In this circuit Q1 and its double-tuned drain circuit comprise a low-gain bandpass amplifier that removes the remaining local oscillator signal while setting the bandwidth at about 50 kHz. At this point, the signal is split into the two channels. In the main channel, Q2, an emitter follower, drives the 50-ohm lowpass delay network. The output from this network passes through the gate (DBM) on to the remainder of the receiver. This particular delay network is a seven-pole Butterworth lowpass filter with a 700-kHz cutoff frequency. The phase shift is about 200 degrees at 500 kHz; therefore, the delay is about 1.1 $\mu$s.

In the noise channel, U2, a MC1590 operating as a video amplifier, is the noise amplifier. It drives Q3, the pulse detector, and CR1, the agc detector. The agc time-constant, set by R1 and C1, is long enough to be unaffected by short noise pulses but will follow the average signal level. The anode of CR1 is biased at one half the supply voltage by R2 and R3. An operational amplifier, U1, amplifies the agc and controls the gain of U2. R4 applies an offset voltage to the input of U1. This has the effect of setting the point at which CR1 begins to conduct, since both inputs of the op amp are at the same potential. R4, therefore, becomes the threshold adjustment. Once set, it should not require further adjustment unless it is necessary to disable the blanker in the presence of a strong adjacent channel signal.

Detection takes place in Q3 and the resulting positive pulses are applied to buffer U3A. The output of U3A triggers the one-shot comprised of R5, R6, C2, C3, CR2, and gates U3B and U3C. The remaining gates of U3 are used to develop the proper phase and current amplitude to operate the blanking gate.

## circuit performance

Figs. 5, 6 and 7 demonstrate the performance that can be achieved with the circuit of fig. 3. The top trace of fig. 5 shows a signal with a simulated noise spike, of 10 dB greater amplitude. The bottom trace is the same signal at the output of the blanking gate.

Fig. 6, made under the same conditions as above, shows, on the top trace, the detected output of Q3. The middle trace is the output of the one-shot as seen at pin 4 of U3D. The bottom trace is the blanker output. The only change in fig. 7 from fig. 5 was to increase the noise to signal ratio to 40 dB and reduce the top trace vertical sensitivity to 500 mV per division. Almost total elimination of the noise at the output is clearly evident.

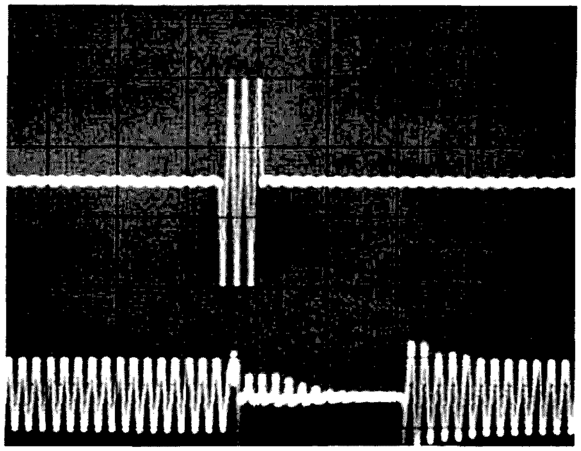

**fig. 7. Example where the noise-to-signal ratio has been increased to 40 dB. The bottom trace shows almost complete elimination of the noise pulse at the output of the blanking gate.**

problems. It is but one technique among other more sophisticated ones, such as coherent detection, adaptive filtering, and auto-correlation, that should be considered when attempting to communicate in the presence of noise.

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**ham radio**

# calculating preamplifier gain from noise-figure measurements

Discussion of a new technique for calculating the gain of vhf/uhf preamplifiers from noise-figure measurements

**Noise-measurement sessions have** become a popular and regular feature of the various vhf/uhf conferences held around the country. In addition to giving the individual experimenter an opportunity to check out his vhf/uhf preamplifiers and converters on precision noise-measuring equipment, the published results allow an annual comparison against the previous year's performance, and provide an index of technical progress in amateur receiver design. In addition, the noise-figure measurements foster a competitive spirit which spurs many experimenters into upgrading their vhf receiver performance. Those amateurs who provide the measuring equipment are similarly inspired to improve the accuracy of their instrumentation and noise-measurement techniques.

The technique presented here, which allows indirect measurement of preamp gain, is an illustration of the latter effect. When I was asked by Wayne Overbeck, N6NB, Chairman of the 1977 West Coast VHF/UHF Conference, to conduct the receiver noise-figure measurements, I was at once flattered and shattered. Flattered because it was an honor afforded to few amateurs — but shattered by the prospects of carting an automatic noise-figure indicator, two noise heads, an i-f strip, five converters, two step attenuators, two power supplies, three signal generators, a power meter, and assorted pads, adapters, and cables 300 miles (480km) each way in the back seat of my small imported car. The most practical suggestion for coping with the latter problem came from one of my students: "Rent a trailer." Not being as practical as he, I had another thought, "Find a way to make the measurements with less test equipment."

## importance of gain information

Fully half the test equipment listed above is used not for measuring noise figure at all, but rather for measuring the gain of the preamplifiers being tested. Gain information is necessary if true preamplifier noise performance is to be evaluated, because to measure the noise figure (NF) of a preamplifier, it must be connected to a converter. The converter will add some noise to the system, and to accurately characterize a preamp, this noise must be subtracted from the measured noise figure. Since the converter's effect on measured noise figure is a function of preamp gain, it is impossible to accurately measure a preamp's noise figure without knowing its gain.

The relationship between preamp gain, converter NF, preamp NF, and the noise figure of the cascade is summarized by Friis' well-known equation[1]

$$F_T = F_1 + \frac{(F_2 - 1)}{G1} \qquad (1)$$

where $F_T$ = total Noise Factor
$F_1$ = preamp Noise Factor
$F_2$ = converter Noise Factor
$G_1$ = preamp gain

all of the above measurements are power ratios (*not* dB).

Since both noise figure and gain are generally expressed in dB, it is necessary to convert to ratios before applying the above formula, thus:

$$F\,(ratio) = Antilog\ \frac{NF\,(dB)}{10} \qquad (2)$$

## why measure gain

Unless a preamp's gain has been measured, **eq. 1** results in two unknowns; to eliminate gain measurements (and avoid renting a trailer), it was evident that

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I had to find a way to solve for both unknowns. My math students reminded me that this could be accomplished only with two simultaneous equations. Clearly what was needed was yet *another* expression for preamp performance which contained no parameters other than those which could be measured on a noise-figure meter.

Inspiration struck when I considered solving Friis' equation for isolating the noise factor of a preamp of known gain, in front of a known converter, from the noise measurement of the cascade:

$$F_1 = F_T - \frac{(F_2 - 1)}{G} \tag{3}$$

Imagine what the expression would be if the *same* preamp were measured in front of *another* converter of noise factor, $F_3$, yielding a new cascade noise factor, $F_{T'}$. Now

$$F_1 = F_{T'} - \frac{(F_3 - 1)}{G} \tag{4}$$

Since $F_T$, $F_{T'}$, $F_2$, and $F_3$ can all be measured on a noise-figure indicator, **eqs. 3** and **4** represent two equations in two unknowns! This means we can now calculate gain solely as a function of noise-figure measurements; no gain measurements are required.

$$F_T - \frac{(F_2 - 1)}{G} = F_{T'} - \frac{(F_3 - 1)}{G}$$

$$F_T - F_{T'} = \frac{(F_2 - 1)}{G} - \frac{(F_3 - 1)}{G}$$

$$G = \frac{(F_2 - 1) - (F_3 - 1)}{F_T - F_{T'}}$$

$$G = \frac{F_2 - F_3}{F_T - F_{T'}} \tag{5}$$

And knowing gain, we can solve for corrected preamp noise factor using either **eq. 3** or **eq. 4**.*

## eliminating the second converter

From **eq. 5** it can be seen that preamp gain data can be derived from two different cascade NF measurements ($F_T$ and $F_{T'}$) of the preamplifier in front of two converters of known and different noise factors ($F_2$ and $F_3$). Although we have eliminated the requirement for signal generators, step attenuators, power meters, and any other equipment associated with gain measurement, we now require two different instrumentation converters for each band at which preamps are to be measured. Hello again, *U-Haul*!

Actually, a single converter for each band can be used for both measurements, by preceeding it with a precision pad for the second measurement to degrade its noise figure by a known amount. $NF_3$ (in dB) would be *approximately* equal to $NF_2$ plus the loss of the pad, both in dB.* Of course, these numbers would be converted to ratios using **eq. 2**, before calculating gain from **eq. 5**.

If a precision 10 dB pad is used to transform $F_2$ to $F_3$, the gain formula becomes:

$$G = \frac{9F_2}{F_{T'} - F_T} \tag{6}$$

The measurement procedure is relatively straight forward:

**1.** Measure $NF_2$, noise figure of the converter

**2.** Measure $NF_T$, noise figure of the preamp-converter cascade

**3.** Insert a precision 10 dB pad between the preamp and converter

**4.** Measure $NF_{T'}$, noise figure of the preamp-pad-converter cascade

**5.** Convert the three NF measurements to power ratios using **eq. 2**

**6.** Calculate gain (ratio) from **eq. 6**

**7.** Calculate corrected preamp noise factor (ratio) from **eq. 3**

**8.** If desired, convert $F_1$ to dB:

$$NF_1\ (dB) = 10\ Log_{10}\ F_1(ratio)$$

## accuracy limitations

Contrary to popular belief, the process of measuring uhf receiver noise figure is highly imprecise, even with such precision equipment as the Hewlett-Packard 340 Automatic Noise-Figure Indicator with argon-discharge noise head. In the near future, Bob Stein, W6NBI, has promised to present a discussion of noise figure indicators and their relative accuracy

*An HP-25 program for calculating noise figure and gain from these equations is available from the author upon receipt of a self-addressed, stamped envelope.

*In fact, the converter-pad combination exhibits a noise figure slightly greater than the sum of pad loss and converter noise figure. This is because there are really two sources of NF degradation when the pad is inserted: the power loss of the pad (its marked attenuation value), and thermal noise generated within the resistance of the pad as a function of its being a warm body (relative to absolute zero). If the attenuation of the pad is fairly high (I use 10 dB), the thermal noise contribution becomes negligible and can be omitted from calculations.

in *ham radio*. Without listing all the various error sources here, suffice it to say that measurements of the type traditionally taken at the regional vhf/uhf conferences are accurate to within only about ± 1 dB.

Why then, do we bother with this annual ritual? Primarily because the measurements made at these conferences are a good *relative* indication of the comparative performance of various designs and devices. You may be confident that the 1296-MHz preamplifier yielding the lowest noise figure at a *particular* competition is indeed the lowest noise-figure device, although, of course, the actual numbers are of limited significance.

I must caution participants in noise-figure competitions against drawing firm inferences from the comparison of data taken on different occasions, on different equipment, or at different times. The fact that all two-meter converters measured this year had lower noise figures than those measured last year is *not necessarily* an indication of technological progress. It's quite possible that differences from year to year are merely a function of divergent measurement errors.

Nonetheless, the "tweak and optimize" procedure generally followed at these noise-figure measurement competitions is entirely valid because the measurements are a good relative indication of receiver performance. I should point out that extensive tuning is not noted for greatly improving noise performance. If the equipment under test is operating reasonably well on the air a minor tweak of the input circuitry, as well as a possible adjustment to bias level, will usually suffice. In fact, I have seen converters tuned to within an inch of their lives, only to end up delivering a much lower NF at some frequency *out of the band* (usually the image frequency).

Along the same lines, note that if tuning the converter's local-oscillator chain results in an indicated NF improvement, it is only because the spurious components generated by optimizing the LO result in multiple mixing products. In short, tweak sparingly!

Since preamp gain information is to be extracted from noise figure measurements, the numbers derived for gain are subject to the same ambiguity which surrounds the measurement of noise figure. Further, considerable measurement error can result if the preamplifier gain should change between the measurement of $F_T$ and $F_{T'}$. Since the input impedance of all converters is not necessarily 50 + j 0 ohms, it is likely that a preamp will see different load impedances with the 10 dB pad installed and removed; if the preamplifier's stability is marginal, the result may be a several dB gain variation. This will adversely effect both gain and NF measurement accuracy, but can be minimized by placing yet another loss pad (aside from the one used to establish $F_T'$) in front of the converter for *both* measurements, to mask input mismatch.

If preamp gain and converter NF remain constant throughout the measurement sequence, this method appears capable of estimating preamp gain to an accuracy on the order of ± 1 dB per 10 dB of gain.

## Santa Barbara field trial

I left my gain measuring gear at home and tried this technique at the West Coast VHF/UHF Conference in Santa Barbara in May, 1977. The results of measuring gain and noise figure of 57 different preamps in the 144, 220, 432, 1296, and 2304 MHz bands correlated closely both with theory and expectations. Several preamps registered unusually high gain, but errors were within the accuracy limits outlined previously. The only severe difficulty encountered was in measuring extremely high-gain (30 dB or so) multi-stage preamplifiers; these tended to overdrive the converters, sometimes introducing enough measurement error to yield values for $F_{T'}$ *lower* than $F_T$! Needless to say, under these conditions the computations fall apart.

Most of the participants in the NF competition were reasonably satisfied with the NF and gain measurements derived from this technique. There were, of course, a few who said, "Your measurements are all screwed up — I *know* my preamp's better than that," but I hear this at all NF competitions, regardless of the equipment or techniques which are used. The method will probably be retained at future West Coast Conferences, and is being recommended for use at the Eastern and Central States events as well.

## disclaimer

I make no claim whatever that the technique presented here for noise-figure measurements is original. However, I have never personally seen the technique applied before, and have no knowledge of anyone else either advocating or using it. But the measurement is so simple, the concept so obvious, that I would fully expect someone, somewhere, has thought of it before. That doesn't matter. What counts is that we hams have yet another measurement tool at our disposal, one which hopefully will enable us to upgrade our vhf and uhf receivers and skills. Please don't consider these measurements sacred; this is, after all, *amateur* radio.

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**ham radio**

# effects of noise in receiving systems

A discussion of the many types of noise which affect communications receivers, including external noise, noisy local oscillators, and noise IMD

**Noise and interference** (man-made noise) as emissions from electrical systems are two limiting factors that determine the full operation of all radio communications equipment. Before we get into a discussion of the effects of noise, however, we must differentiate between the different sources of radio noise — noise effects can be broken down into four general categories:

**1.** Atmospheric noise, precipitation static, galactic noise, and man-made noise.

**2.** Noise performance of the rf and i-f amplifier stages which, together with mixer losses, determine the overall noise figure of the receiver.[1]

**3.** Noisy local oscillators which cause problems with blocking and reciprocal mixing.

**4.** Noise intermodulation distortion which occurs as in-band and out-of-band products.

Since the noise sources listed in category 1 have been widely discussed in the past,[2] they will be mentioned only briefly here. The main thrust of this article will be in categories 2, 3, and 4: noise performance of rf and i-f amplifier stages, noisy local oscillators, and noise IMD.

## external noise

Atmospheric noise is produced primarily by lightning discharges associated with thunderstorms, so the level of atmospheric noise depends upon frequency, season of the year, time of day, and geographical location. Graphs similar to that shown in **fig. 1**, as well as plots of noise field strength such as that shown in **fig. 2**, are available from the National Bureau of Standards. These can be used to determine received signal strength and signal-to-noise ratio, once the frequency, distance, and time of day are chosen.

**Fig. 3** shows the optimum frequency for communications over a distance of 20 miles (32km). This particular plot is for a manpack radio application, but similar charts could be prepared for amateur radio communications. In this case 3.6 MHz is the op-

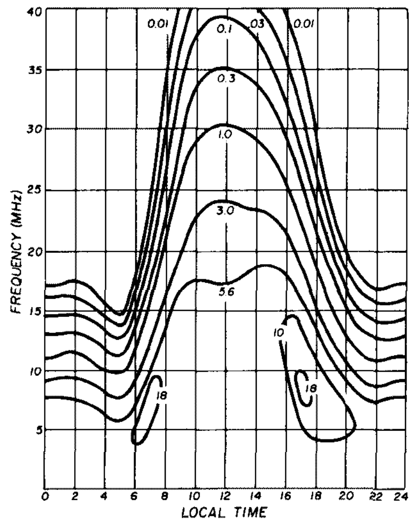


**fig. 1. Field strength in microvolts per meter for a 1000-watt transmitter as a function of local time and frequency.**


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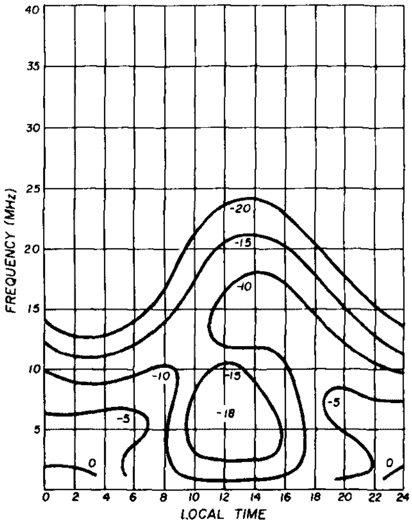



**fig. 2. Atmospheric noise field strength in dB above or below 1 $\mu$V per meter as a function of time and operating frequency (assumes 1 kHz bandwidth). If this noise chart is placed over the propagation chart in fig. 1, the signal-to-noise ratio can be calculated (S/N = receiving field strength minus noise field strength).**

timum frequency until slightly after 0500 (point 1), when 6.3 MHz is the better choice. At 1600 the operator changes frequency to 8.2 MHz (point 2), and at 2000 goes back to 6.3 MHz (point 3). At about 2300 hours local time, 3.6 MHz is again the optimum frequency (point 4) and remains so until the next morning at about 0500.

Precipitation static is produced by wind and rain around the receiving antenna, and is important only below about 10 MHz; it is particularly troublesome for aircraft communications. It can be reduced by

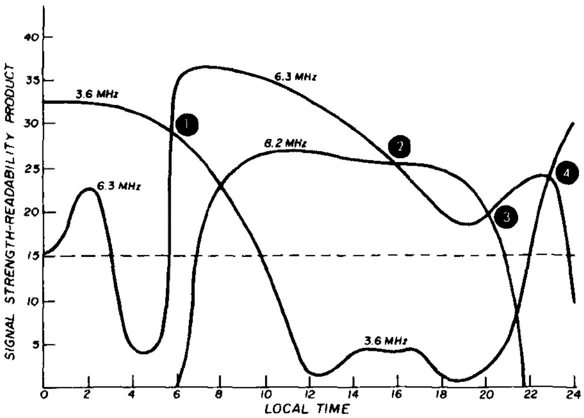



**fig. 3. Optimum operating frequencies vs time of day for two man-pack radios operating over a distance of 20 miles (32km). This data was gathered in September, 1969. The dashed line indicates the minimum required signal strength-readability product to understand the transmitting station.**

providing a dc dissipation path for any charge which may build up on the antenna.

Galactic noise (also called cosmic noise) is defined as rf noise which is caused by disturbances which originate outside the earth or its atmosphere. The primary causes of this noise, which extends well into the microwave region, are the sun and a large number of radio sources distributed along the Milky Way.

Man-made noise is generated by rotating electrical machinery and automobile ignition systems, among other things, and is the predominate external noise source if you live in (or near) an urban area.

## definition of noise

The level of electrical noise is most conveniently referred to as noise power (assuming wideband noise):

$$P_n = kT_oB \qquad (1)$$

where $P_n$ = available noise power

$k$ = Boltzmann's constant
= $1.38 \times 10^{-23}$ joules/°Kelvin

$T_o$ = reference temperature
(typically 290°K)

$B$ = effective receiver noise bandwidth, Hz

This can also be written in terms of the root-mean-square noise voltage

$$e_n^2 = 4kT_oBR \qquad (2)$$

where $e_n^2$ is the open-circuit root-mean-square noise voltage, and $R$ is the resistance (see **fig. 4A**). Since maximum signal will be transferred to a load when the load resistance is matched to the source resistance (**fig. 4B**), the root-mean-square voltage across the load, when it's impedance matched to the source, is given by

$$e_n^2 = 4kT_oB\frac{R}{2} \qquad (3)$$

For example, if a receiver front end is matched to a 50-ohm resistor, and the noise bandwidth, $B$, of the receiver is 2.4 kHz, calculate the noise voltage across the antenna terminals of the receiver at room temperature ($T_o = 290°K$).

$$e_n = \sqrt{4(1.38 \times 10^{-23})(290)(2.4 \times 10^3)(50/2)}$$
$$= 3.1 \times 10^{-8} \text{ volts}$$
$$= 31 \text{ nanovolts (nV)}$$

When you're dealing with wideband communications systems where the noise bandwidth may be unknown, it is more convenient to work with noise factors or noise figures, rather than noise power or noise voltages.

In estimating the noise at the receiving system due to external sources, it must be remembered that noise power is proportional to the bandwidth, $B$,

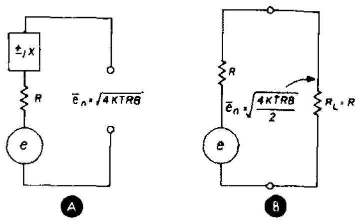



**fig. 4. Mean noise voltage depends on temperature, resistance, and noise bandwidth (A); this is the open-circuit noise voltage. Maximum noise power is transferred to the load when the load resistance is matched to the source resistance (B); in the terminated condition only half the available noise power appears across the load (receiver) terminals.**

assuming uniform white noise. The receiver noise factor is given by

$$F = \frac{P_n}{kT_o}\ B = \frac{T_a}{T_o} \tag{4}$$

where $P_n$ is the available noise power and $T_a$ is the effective noise temperature (°K). If the noise factor

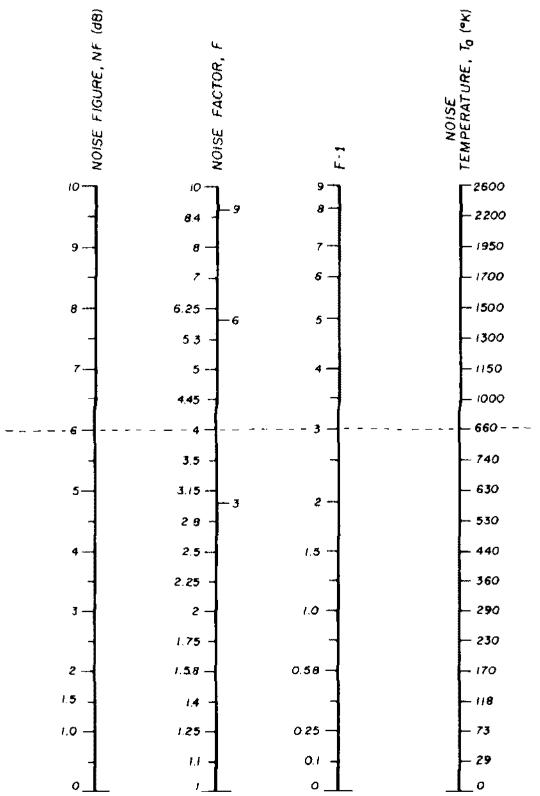



**fig. 5. Conversion table for noise figure, noise factor, and effective noise temperature. The dashed line shows a noise factor of 4 is equivalent to a noise figure of 6 dB or an effective noise temperature of 860°K.**

of a receiving system is known, the noise voltage for a signal-to-noise ratio of unity ($S/N = 1$) can be calculated from

$$e_n^2 = F(4kT_o\,BR) \tag{5}$$

For a noise factor of 10 (noise figure, $NF = 10$ dB), 2.4 kHz bandwidth, and $R = 50/2$

$$e_n = \sqrt{10 \cdot 4(1.38x10^{-23})(290)(2.4x10^3)(50/2)} = 98\,nV$$

Since most receiver specifications are written for a 10 dB signal-to-noise ratio, the terminated input voltage for this ratio is 310 nV or 0.31 $\mu$V.

Note that this calculation is made with noise factor, $F$, *not* noise figure $NF$. The two must not be confused. Noise figure is simply

$$Noise\ figure = 10\ log\ noise\ factor$$

A simple graph for converting from noise factor, $F$, to noise figure in dB or effective noise temperature, $T_a$, is presented in **fig. 5**. If you wish to specify antenna noise as a function of frequency, noise temperature is the most convenient way to do this, as shown in **fig. 6**.

The overall noise factor of a series of amplifier stages connected in cascade is given by

$$F = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \frac{F_4 - 1}{G_1 G_2 G_3} + \ldots + \frac{F_n - 1}{G_1 G_2 G_3 \ldots G_{n-1}} \tag{6}$$

where $F_1$ and $G_1$ are the noise factor and available power gain of the first stage; $F_2$ and $G_2$ are those of the second stage; $F_3$ and $G_3$ are those of the third

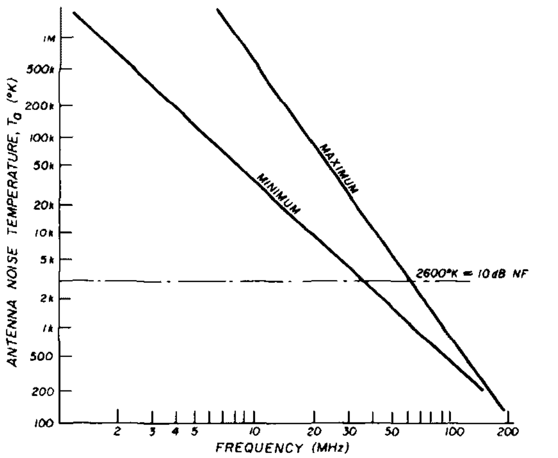



**fig. 6. Noise temperature of an antenna (unity gain) as a function of frequency under daytime and nighttime conditions. The dashed line shows that a receiver noise figure of 10 dB is sufficient up to about 38 MHz.**

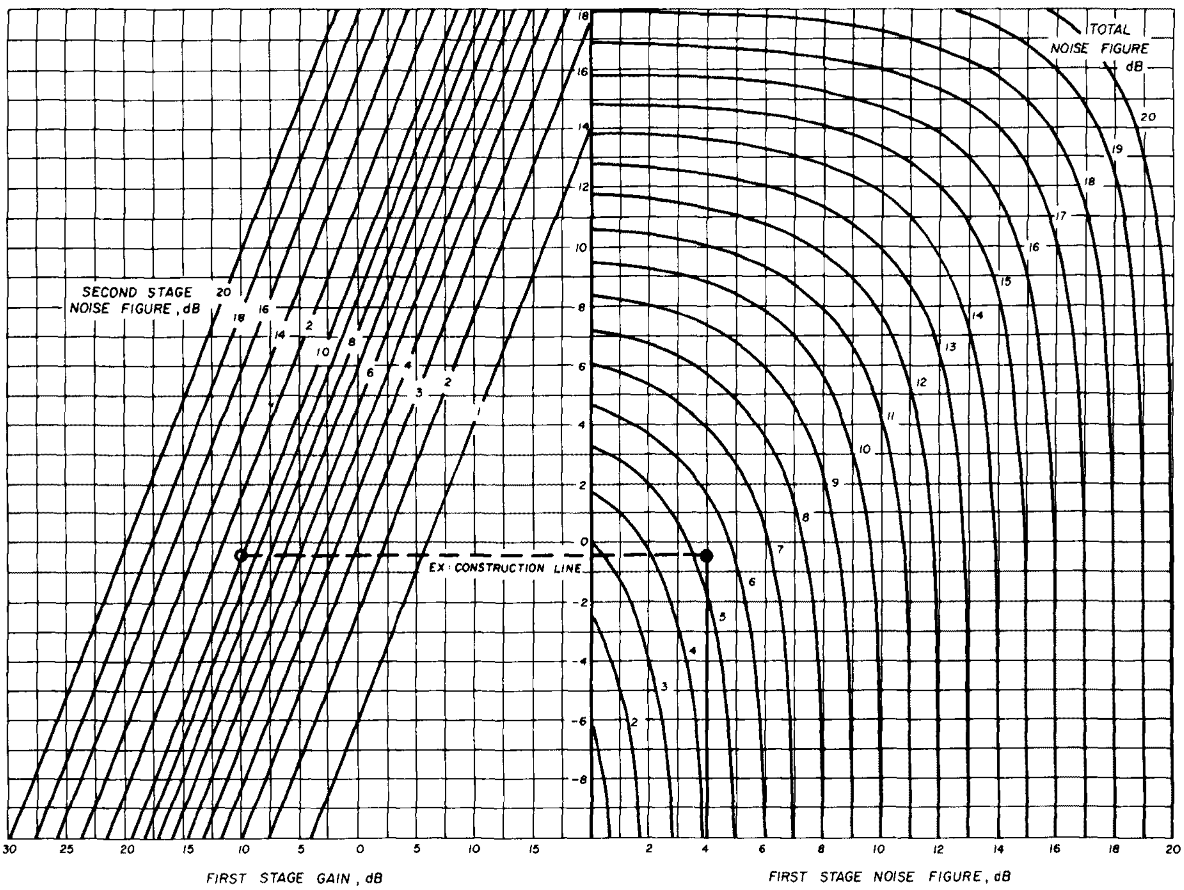



**fig. 7. Chart for calculating the overall noise figure of two cascaded stages. In the example shown, a first stage with a noise figure of 4 dB and power gain of 10 dB, is cascaded with a stage having a 10 dB noise figure; the overall noise figure is 5.3 dB.**

stage; $F_n$ is the noise factor of the nth (last) stage, and $G_{n-1}$ is the gain of the next to last stage. For the case of two amplifier stages, this equation can be simplified to

$$F = F_1 + \frac{F_2 - 1}{G_1} \tag{7}$$

**Fig. 7** is a graphic solution to this formula for cascaded noise figure, in dB, for two stages. In the example shown in the chart, the first stage has a noise figure of 4 dB and a power gain of 10 dB; the second stage has a noise figure of 10 dB. The overall noise figure of the two cascaded stages is 5.3 dB. This is discussed further in reference 3.

## amplifier noise

The noise performance of rf and i-f amplifiers is dictated by the transistors used in the circuit, and the selection of optimum operating points. Sometimes there are tradeoffs between low noise or low distortion; it is up to the circuit designer to decide which is more important. The noise figures of transistors and ICs are specified by the manufacturers, and while field-effect transistors generally have lower noise

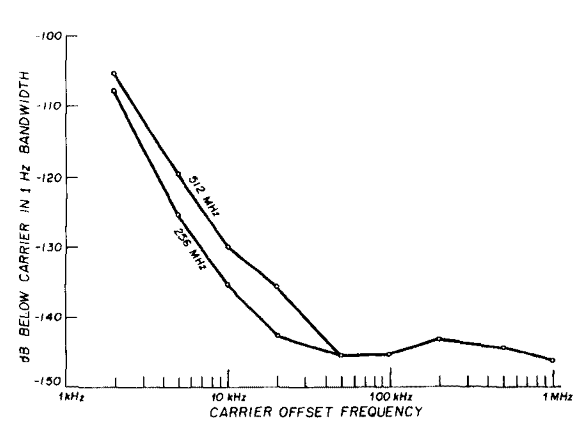



**fig. 8. Noise sideband performance of the Hewlett-Packard 8640B signal generator.**

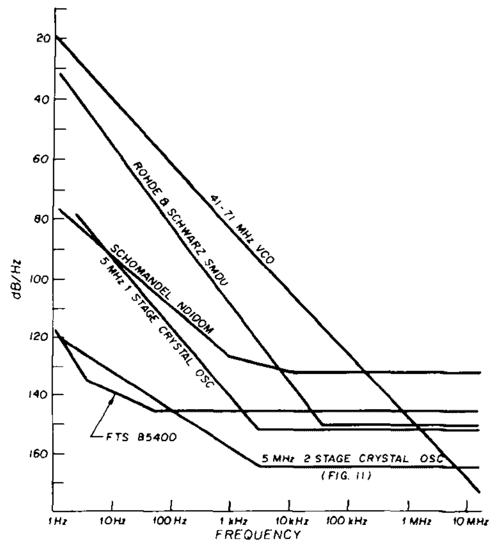



**fig. 9. Measured noise sideband performance of a 41-71 MHz vco, Rohde & Schwarz SMDU signal generator, Schomandel ND100M frequency synthesizer. Frequency and Time Services (FTS) B5400 modular 5-MHz crystal oscillator, and single and double stage 5-MHz crystal oscillators.**

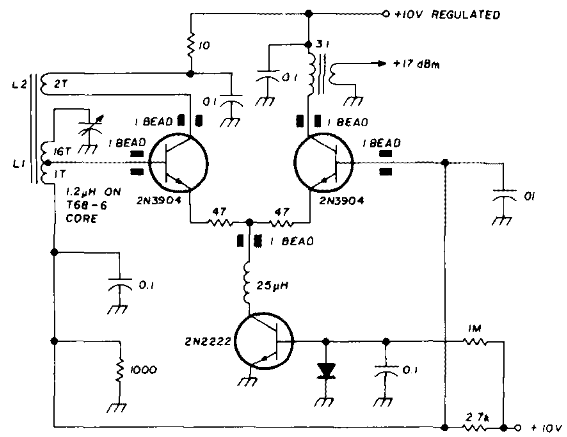



**fig. 10. Schematic of a very low noise LC oscillator operating at 5 MHz.**

figures, their second-order distortion is much worse (by definition) than that of bipolar transistors (the square-law characteristic of the fet makes it ideally suited as a frequency doubler). Because of their high gain reserve, it's relatively easy to linearize bipolar transistors with suitable feedback circuits (transistors offer higher gain-bandwidth products than fets with slightly higher noise figures).

## noise in oscillators

The graph in **fig. 8** shows the sideband noise of an oscillator stage using a high-$Q$, tuned LC circuit. To analyze the noise performance of the oscillator let's first assume that the oscillator does not have its feedback loop closed and operates as a linear amplifier. It has been proven experimentally that noise modulation, which results in so-called phase noise generated in rf amplifiers as a modulation of the carrier, will produce close-in noise of 115 dB/Hz. The worst and best cases are 100 and 120 dB/Hz, depending upon the transistor used in the circuit. This is especially true of noise 1 kHz to 5 kHz from the carrier, and is caused by flicker noise. In accordance with reference 3, close-in oscillator noise can be improved only through the use of negative rf feedback (emitter degeneration); I described several circuits using this technique in reference 4. The use of rf negative feedback allows a signal-to-noise improvement of 40 dB, which will result in 150 dB/Hz sideband noise close-in to the carrier.

To achieve the ultimate in signal-to-noise ratio, amplifiers which are driven by only one carrier require heavy feedback; they should be operated class A or class AB1 to avoid noise sideband intermodulation.

Noise sideband measurements have always been a gray area, and many publications have stated incorrect figures or shown bad circuits. As a general rule, when designing oscillators, use the following guidelines:

1. Oscillators should always use two stages: one operating in class A, and the other operating as a limiter. The limiter is also used as the feedback part (the intermodulation distortion introduced by the limiter is partially improved by the feedback loop).

2. Circuits which have agc applied to the oscillator transistor should be avoided because the agc will likely add noise. This is discussed in detail in reference 6.

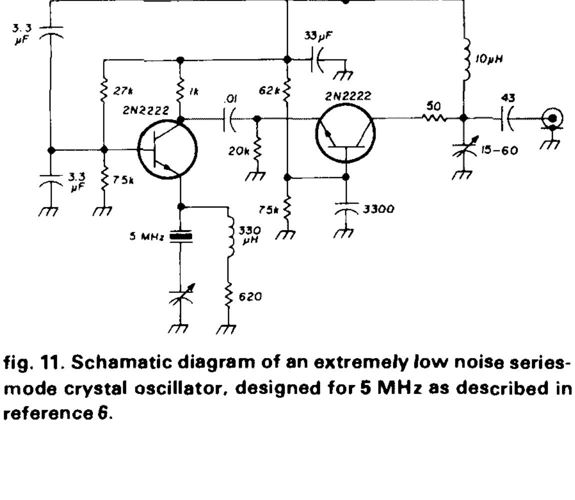



**fig. 11. Schamatic diagram of an extremely low noise series-mode crystal oscillator, designed for 5 MHz as described in reference 6.**

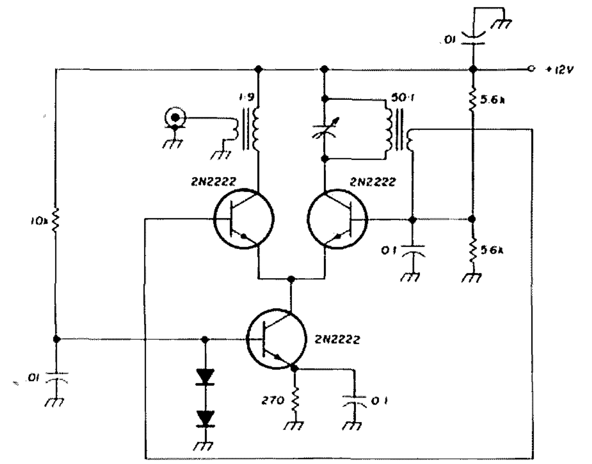



**fig. 12. Circuit for a low noise and low distortion LC oscillator. Harmonic distortion is less than 1 per cent.**

3. Statements about the use of ECL integrated circuits as low-noise oscillators are basically wrong. In addition, the noise performance of field-effect transistor oscillators is not necessarily better by definition; feedback techniques must be used for good low-noise performance.

**Fig. 9** shows the measured noise sideband performance of several signal generators, synthesizers, and oscillators. **Fig. 10** shows a very low noise LC oscillator circuit, while **fig. 11** shows a very low noise crystal oscillator circuit which can be used with both fundamental and overtone crystals.

In some applications it is desirable to have an oscillator circuit with low harmonic distortion — this can be accomplished with a differential oscillator circuit which combines low noise with low harmonic distortion. **Fig. 12** shows an oscillator with low harmonic distortion output.

Crystal oscillator circuits which are used in frequency synthesizers are usually optimized for good aging performance and do not have the lowest noise figures. The noise figure of these oscillators can be improved by adding a crystal filter with less than 20 Hz bandwidth (the crystal filter is sometimes incorporated in the same proportional oven as the reference crystal). **Fig. 13** shows the schematic of such a crystal filter. The bandwidth of a simple crystal filter such as that shown here is determined by the input and output impedance; the series resonant resistance, $R_s$, which determines the $Q$ of the crystal, is increased by the series connection of the input and output impedance, which effectively lowers the $Q$ of the crystal. Therefore, when building crystal oscillators for low-noise applications, it is vital that the drive and load impedance are equal or less than the series resonant resistance. The degradation of $Q$ results from the relationship

$$\frac{Q_o}{Q_L} = \frac{R_s}{R_s + R_{T1} + R_{T2}} \qquad (8)$$

where $Q_o$ = unloaded $Q$
$Q_L$ = loaded $Q$
$R_s$ = series resonant resistance
$R_{T1}, R_{T2}$ = differential terminating resistors

In the circuit of **fig. 13**, this is accomplished by setting the collector current to 15 mA.

## noise in receiving systems

Both external noise and oscillator noise have been

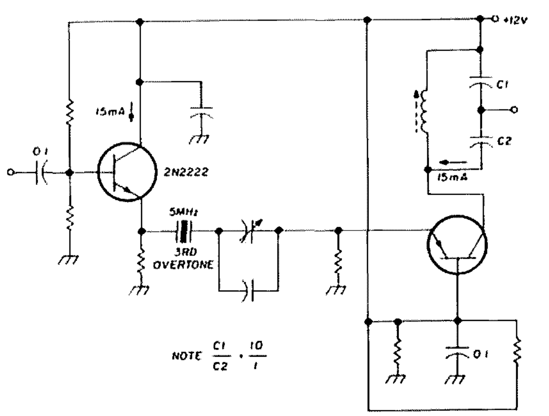



**fig. 13. Schematic of a simple crystal filter with only a few Hz bandwidth. Collector current is set at 15 mA to satisfy the relationship between the series-resonant resistance and unloaded Q, as discussed in the test.**

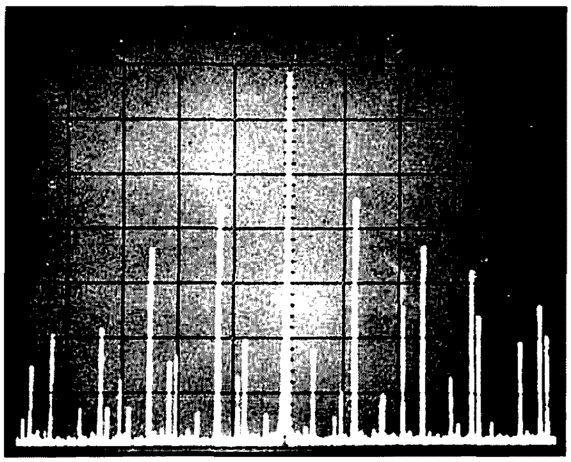


**fig. 14. Output of a double-balanced mixer with one strong station and several weaker signals applied. This produces numerous close-in spurs or birdies. The same thing occurs when the LO drive is insufficient.**

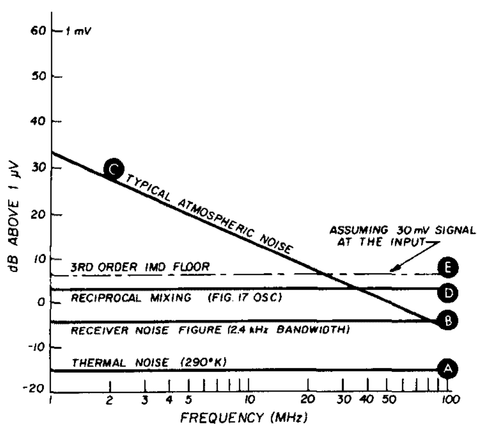



**fig. 15. Typical energy distribution. Thermal noise = 31 nV (A), receiver noise = 95 nV (B), typical atmospheric noise (C), effect of reciprocal mixing (D), and third-order IMD (E).**

discussed, but there is still another source of noise to be considered. Certain intermodulation distortion products which are generated in the mixer, and are either produced from overloading the mixer or are reversed modulation of the oscillator sideband noise, must be considered in a recceiving system.

Fig. 14 is a spectrum display of the output of a mixer with various frequencies fed to the input. If the mixer has a third-order intercept point of +30 dBm, this means that when two tones, each 0 dBm, spaced $\Delta F$ apart, are applied to the rf port, two mixing products, $\Delta F$ above and below the two input signals, will be generated at a level of − 60 dBm or 224 μV.

If there are conditions of good radio propagation, the output of a full-size, wideband rhombic antenna will be between 30 and 100 mV. If we assume for a moment that there is a quasi-infinite number of stations operating between 9 and 12 MHz, spaced 5 kHz apart, then we will also have an intermodulation distortion product every 5 kHz, which will be 72 to 100 dB below the input signals. Under the worst conditions, this will produce an intermodulation distortion floor of − 72 dBm or 60 μV. Under these conditions no signals below 60 μV can be detected.

Fortunately, on the average these spurious products are not closer than about 5 to 6 kHz. For ssb or a-m reception this means that a narrowband crystal filter immediately following the first mixer will cure the problem since all other third-order and higher products are outside the passband of the crystal filter. This simple solution wouldn't work if we had to deal with CW stations a few 100 kHz apart, but this is not usually the case. And it should be remembered that only broadcast, certain marine, and some point-to-point stations generate enough radiated power to create such large input signals.

When analyzing the shortwave broadcast bands, it's interesting to compare received signal strengths in America with those in Europe. The Eastern-bloc countries, which are transmitting with excessive power, are 20 dB stronger in Europe than they are here, so the design requirements for a short-wave receiver for commercial applications in the United States are somewhat less than they would be for the same application in Europe.

Going back to the question of CW stations, the single-conversion receiver, which uses the lowest possible bandwidth immediately following the first mixer, will practically always outperform a double-conversion receiver. This can be checked easily by comparing the single-conversion receiver in the Drake TR4C in the CW mode with the Drake R4C, which uses double conversion. During crowded CW conditions, such as a CW contest, some agc pumping occurs in the R4C. The only way to prevent this from happening is to keep the gain between the first and second mixer as low as possible, and to use a second mixer with the same basic intercept point as the first mixer.

In a recently published paper, the chief engineer of the Racal Company in England nicely demonstrated the response of double conversion and the influence of blocking or reciprocal mixing.[7] While the idea of so many unwanted mixer products, even at low input levels, may be shocking, it must be remembered that the man-made noise splatter due to overmodulation of a-m and ssb transmitters, splatter because of high-speed CW, and other radiated rf energy, will fall above the receiver's sensitivity. **Fig. 15** shows the energy distribution for five different interference

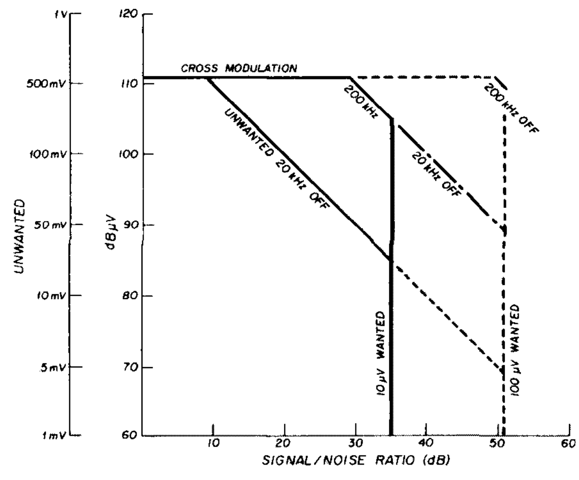
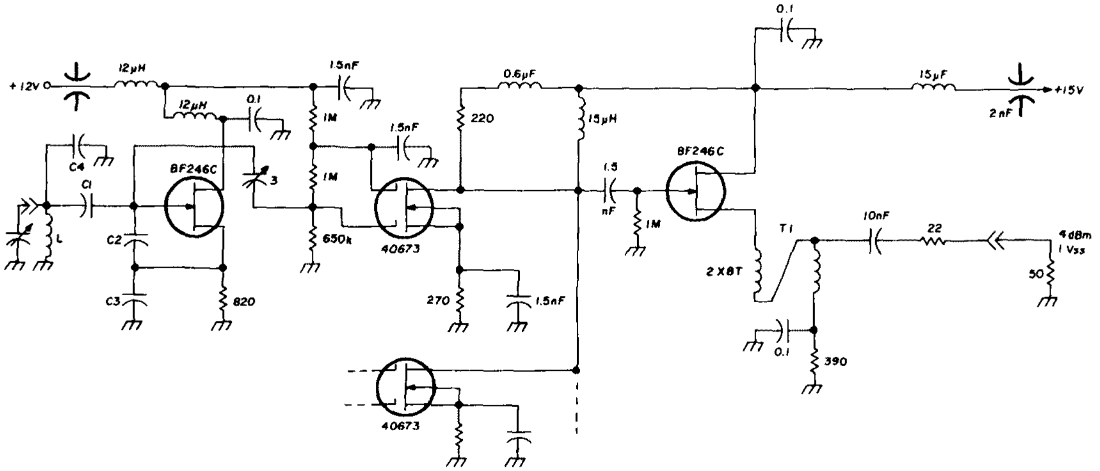



**fig. 16. Graph showing the effect of reciprocal mixing and cross modulation, assuming a wanted signal of 10 μV, and one of 100 μV.**




| band | MHz | C1<br>pF | C2<br>pF | C3<br>pF | C4<br>pF | L<br>μH | L-Typ Nr.<br>*Fa. Stettner* | ΔC<br>pF |
|---|---|---|---|---|---|---|---|---|
| 80 | 11,963-13,076 | 68NPO<br>15P100<br>15P100 | 100NPO<br>250N150<br>15P100<br>15P100 | 680N150 | 0 | 1,9P30 | 87-5319/47 | 10-30 |
| 40 | 15,948-17,049 | 68NPO<br>15P100<br>15P100 | 100NPO<br>250N150 | 470N150 | 1,5NPO<br>2N150 | 1,1P30 | 87-6023-01 | 10-26 |
| 20 | 22,951-24,083 | 56NPO | 100NPO<br>39N150<br>15N220 | 330N150 | 2N150 | 1,0P30 | 87-5319/35 | 5-12 |
| 15 | 29,950-31,060 | 100NPO<br>15N220 | 100NPO<br>15P100<br>15P100 | 330N150 | 5P100 | 0,4P30 | 87-5880e-01 | 5-18 |
| 10 | 36,980-39,010 | 56NPO | 100NPO<br>30N470<br>30N470 | 250N150 | 1-5NPO | 0,32P30 | 87-5880d-01 | 5-12 |

**fig. 17. Circuit for a low-noise, fiva-band LC oscillator for amateur equipment (designed by Michael Martin, DJ7VY). The temperature coefficient of each capacitor is given (P = positive coefficient; N = negative temperature coefficient).**

sources which can affect the ultimate sensitivity of a receiver:

**A.** Thermal noise (290°K, − 174 dBm = 31 nV)

**B.** Receiver noise (2.4 kHz bandwidth, 95 nV)

**C.** Atmospheric noise as a function of frequency (typical)

**D.** Affect of reciprocal mixing (see reference 8)

**E.** Third-order IMD products assuming an infinite number of stations.

**Fig. 16** shows the effect of reciprocal mixing and cross modulation with wanted signals of 10 μV and 100 μV. It can be clearly seen that reciprocal mixing occurs long before cross modulation saturates the receiver.

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**ham radio**

# second thoughts on the direct-conversion receiver

Stage-by-stage account of direct-conversion receiver design — a real Dutch treat for receiver buffs

**After some false starts** the direct-conversion (DC) receiver now seems to be well established in amateur radio along with the classic superheterodyne. Simplicity is no doubt the main feature of the DC receiver. Compared with the superhet it lacks an i-f amplifier and second detector, but it has some assets that make it a very fine receiver indeed.

In this article we'll review the basic elements of the direct-conversion receiver as well as some refinements that can be added. Strong and weak points of the design are discussed. Also presented are some ideas you can use should you wish to build a DC receiver to suit your own needs.

## design principle

The DC receiver basic design is shown in **fig. 1**. The signal from the antenna enters the mixer after preselection by L1, C1. An rf amplifier could be included, but it is not necessary in most cases. In the mixer the signal is heterodyned with the signal from the local oscillator (vfo) — exactly as in the superhet — the frequency of which is determined by tuned circuit L2, C2. For ssb reception the vfo is tuned to the frequency of the suppressed carrier. For CW the vfo is detuned as many Hz as the pitch of the note you want to hear.

The DC receiver is not suitable for reception of amplitude-modulated signals. In practice, a-m speech can be heard by tuning zero-beat with the carrier, but this is not an elegant solution. Neither can fm be received. RTTY may be received, provided it is transmitted as fsk, not as a-m.

The mixer is followed by an af filter, which sets the selectivity. Filter bandwidth can be different for ssb and CW. An af amplifier increased the signal to the proper level for headphones or speaker.

In the past, receivers were judged mainly on their selectivity and sensitivity. Nowadays, receiver behavior in the presence of strong signals is a major consideration. It is interesting to compare the DC receiver and the superhet with regard to phenomena that can be called "unwanted signals:"

| unwanted signal | superhet | DC receiver |
|---|---|---|
| I-f breakthrough | yes | no |
| Reception on image frequency | yes | no |
| Reception by mixing with oscillator harmonics | yes | yes |
| Crossmodulation | yes | yes |
| Intermodulation | yes | yes |
| Breakthrough of a-m stations outside passband | no | yes |

Because the first two unwanted signals don't exist with the DC receiver, preselection can usually be somewhat simpler than with the superhet (more about this aspect later). Whether or not the DC

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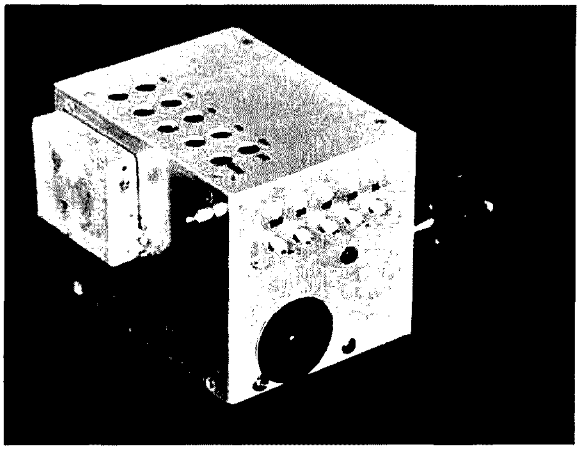
receiver suffers from images can indeed be a matter of opinion. As is well known, the difference in frequency between signal and image frequency amounts to twice the i-f in the case of the superhet. One could consider the DC receiver as a superhet with an i-f of zero hertz. In that case, the frequency difference between wanted signal and image frequency would also be zero. And that is indeed so: a signal 500 Hz *above* the tuned frequency, say, is received just as well as a signal 500 Hz *below* the tuned frequency. But instead of talking about images, one could also state that the DC receiver can't discern between upper and lower sideband.

And here we also have put our finger on one of the weak points of the direct-conversion principle. It is of course possible to incorporate means for suppressing one sideband (phasing), but to my mind this entirely spoils the main attraction — simplicity — of the DC receiver.

Breakthrough of strong a-m stations is typical for the DC receiver. We will thoroughly investigate this matter also. There are more, but less serious, snags: hum on peaking the input tuned circuit and microphonics of this circuit. These will also be dealt with.



**The five-in-one vfo in its shielded box. Small box on left contains the buffer amplifier. The row of holes to the right on top and the tie point at left front are for a receiver incremental tuning control, a feature that has no significance in this case and therefore is not shown in the schematic of fig. 5.**

## mixer

We have already mentioned a-m breakthrough. This phenomenon manifests itself by a multitude of speech and music signals that are independent of vfo tuning. The signals come from extremely strong broadcast stations insufficiently attenuated by tuned circuit L1, C1 in fig. 1. In Europe one finds many of these strong stations between 4 and 8 MHz. Out of curiosity I once tuned the input circuit of my receiver to one of these stations near 4 MHz in the evening. The circuit uses a powdered-iron toroid for the coil. For an antenna I used one-half of my 40-meter long inverted vee transmitting antenna and an open-line feeder 20 feet (6m) long. The feeder was coupled to the tuned circuit by running the wire once through the hole in the toroid. At times a vtvm connected to the top of the otherwise unloaded circuit read 0.7 volt!

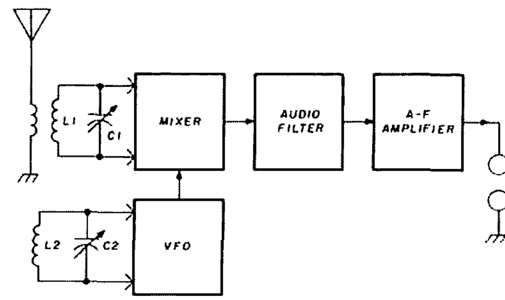



**fig. 1. Block diagram of the direct-conversion receiver.**

The most important factor that determines whether or not a-m breakthrough occurs is the type of mixer used (or product detector, if you prefer). In many designs for DC receivers, a 40673 mosfet or similar device is used as a mixer, probably because an fet has considerable freedom from cross-modulation and intermodulation due to its quadratic characteristic. This benefits both superhet and DC receiver. However, one important aspect is often overlooked: a quadratic characteristic also produces detection of amplitude-modulated signals. Mathematical treatment of a receiver detector almost invariably starts on the assumption of a nonlinear device with a second-power characteristic. That a mixer with a second-power characteristic in a DC receiver *does* detect a-m has been noted by many users to their dismay.

The same problem occurs in a superhet of course, but the resulting audio signals in the mixer output can't pass the i-f amplifier output and thus won't be noticed. But in the DC receiver, no i-f amplifier is present, so every af component out of the mixer, whether from the wanted signal or an unwanted broadcasting station, reaches the output of the receiver.

It is possible to improve the rejection of a-m signals by using two transistors in a balanced mixer configuration. An example can be found in reference 1.

Fig. 2 comes from that article and shows the principle. The difficult part of this circuit is with transformer T1; the circuit should be designed for a high-impedance on the primary and secondary side and have a center tap on the input winding. For CW a tuned circuit can be used, as YU2HL did, but a suitable transformer may be hard to find for ssb.

In my opinion, the only correct solution is to use a mixer that does not depend on a nonlinear characteristic (of second or higher power) but on a switch that opens and closes with the vfo frequency — in other words, a *switching-type demodulator.* A diode alternately brought into conduction and nonconduction by the vfo signal can be used. During transition from one state to the other, a-m detection can still occur on the curved part of the diode characteristic. It is therefore important to make the duration of these transitions as short as possible by driving the diode hard with a strong oscillator signal. Moreover, the antenna signal is also present on the diode, and this signal should not influence the switching characteristic, as this would also cause trouble. Again, this requires an oscillator signal that is strong with respect to the antenna signal. Even if these precautions are taken, some a-m detection will occur when using a single diode.

The remedy is to put two diodes in a balanced-mixer. Several suitable circuits can be found in the literature, but all have one element in common: a potentiometer to set the balance. My experience has been that the setting of this pot depends both on the frequency being used and the oscillator-signal amplitude. This is rather awkward, and in a multiband receiver one must put the pot on the front panel. Another drawback of these single-balanced mixers is considerable conversion loss.

Another possibility is to use a single- or double-balanced mixer using bipolar transistors. I started my experiments with direct conversion using a Plessey SL640 integrated-circuit double-balanced mixer. The balance in such an IC mixer is inherently very good. Nevertheless, I found suppression of a-m breakthrough disappointing. Also the IC produced more noise than I liked, resulting in poor sensitivity.

My experiments led to the conclusion that the only entirely satisfactory mixer for the DC receiver is a double-balanced mixer using four diodes. I tried germanium, silicon, and hot-carrier (Schottky barrier) diodes and homemade input and output transformers, both on ferrite toroid cores and powdered iron cores with two holes. All gave good results. The rf signal at the mixer for a 10-dB signal-plus noise-to-noise ratio was of the order of 3.1 microvolts for the 1.8-21 MHz bands.

Suppression of a-m breakthrough was measured by injecting a 30-percent modulated 400-Hz signal into the receiver and noting the amount of generator amplitude required to cause a 10 −dB s+n/n ratio at the output. (The ssb filter was in operation for this test). The signal was detuned outside the receiver

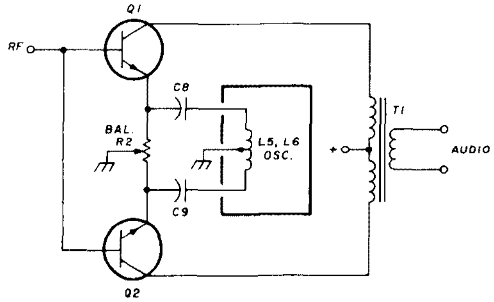



**fig. 2. Balanced transistor mixer used by YU2HL (from reference 1).**

passband so that, without modulation, no output could be measured. An average of about 50 microvolts of the 30-percent modulated signal was needed. In practical terms this meant that, using my inverted-vee antenna with an antenna tuner coupled to the DC receiver, no a-m detection was noted on any band at any time.

The voltage mentioned is the *emf* of a signal generator with 50 ohms output impedance. Note that often the receiver *input signal* is mentioned as producing a certain s+n/n ratio and, for this voltage, half the generator emf is taken. This is correct only when the generator is power matched to the device under test, and in many cases this cannot be relied upon. So it's better to state the generator emf and accept the fact that the sensitivity figures look poorer.

A great surprise came when I replaced the homebrew mixers with an Anzac MD108 double-balanced mixer (the most inexpensive on the Dutch market).* This mixer is specified for the range 5-500 MHz. However, I wanted my receiver to operate on 1.8 and 3.5 MHz as well. Any doubts proved to be unfounded as suppression of a-m breakthrough was about the same as my own mixers. But sensitivity improved to an average of 0.82 microvolt for the bands 1.8 through 21 MHz!

This reveals another very good reason for using a well-balanced mixer: it appears that with my homemade mixers, receiver sensitivity was limited by noise that amplitude modulated the oscillator signal. In a perfect double-balanced mixer this noise is balanced

*And probably the most easily available to American amateurs as well. The MD108 is priced at $7.00 (plus postage) and is available from the manufacturer, Anzac Electronics, 39 Green Street, Waltham, Massachusetts 02154.

out, but my mixers obviously were far from perfect. It is for the very same reason that balanced diode mixers are used in radar receivers without rf amplification. So it's better to buy a good double-balanced mixer right away unless you know the secret of making a good one yourself.

## input circuit

A schematic of my direct-conversion receiver without the vfo is shown in **fig. 3**. Resistor R1 is used to attenuate strong signals. Often this circuit appears as shown in **fig. 4A**. The result is that the tuned circuit becomes more heavily damped and thus less selective as the slider on R1 is moved downward to increase attenuation. Of coourse we would prefer it the other way. This is accomplished by the circuit of **fig. 4B**. Usually R1 is a carbon pot of about 500 ohms. I found that with this control response was uneven over the pot travel; moreover, the pot became defective after some use. Eventually I used a 220-ohm wire-wound pot for R1. Although a wire-wound pot at radio frequencies is completely against the rules, control was good. On higher-frequency bands maximum attenuation became less but was still more than enough.

The task of the input tuned circuit is to suppress products caused by signals mixing with vfo harmonics. If the double-balanced mixer were perfect and the vfo had no even harmonics, reception could occur only on odd harmonics (the third harmonic being the lowest). In practice the receiver shows some sensitivity on the second harmonic as well, although less than on the third. Experience shows that a single-tuned circuit with good loaded *Q* at the input is sufficient to suppress all but the strongest unwanted signals on harmonic frequencies. The one or two signals that remain are easily recognized as they peak at a different setting of C1.

As shown in **fig. 3** I used two separate input tuned circuits: one for 15 through 40 meters and the other for 160 and 80 meters. Each circuit is tuned by one-half of a split-stator capacitor of the type used in broadcast receivers. The advantage is that no switching is necessary within the critical circuits: only input and output links are switched by a good-quality toggle switch. However, a single-variable capacitor could be used that is switched between two or more coils.

No coil-winding details are given for the input circuits as these coils depend on form factor and type of core. A toroid core of ferrite or powdered iron is preferable. L1 and L3 can, as a start, be made equal; the same applies to L4 and L5. If the link coils are small, the circuit will have little damping and selectivity will be good but signal loss will be large. With many turns for the links, little signal will be lost but

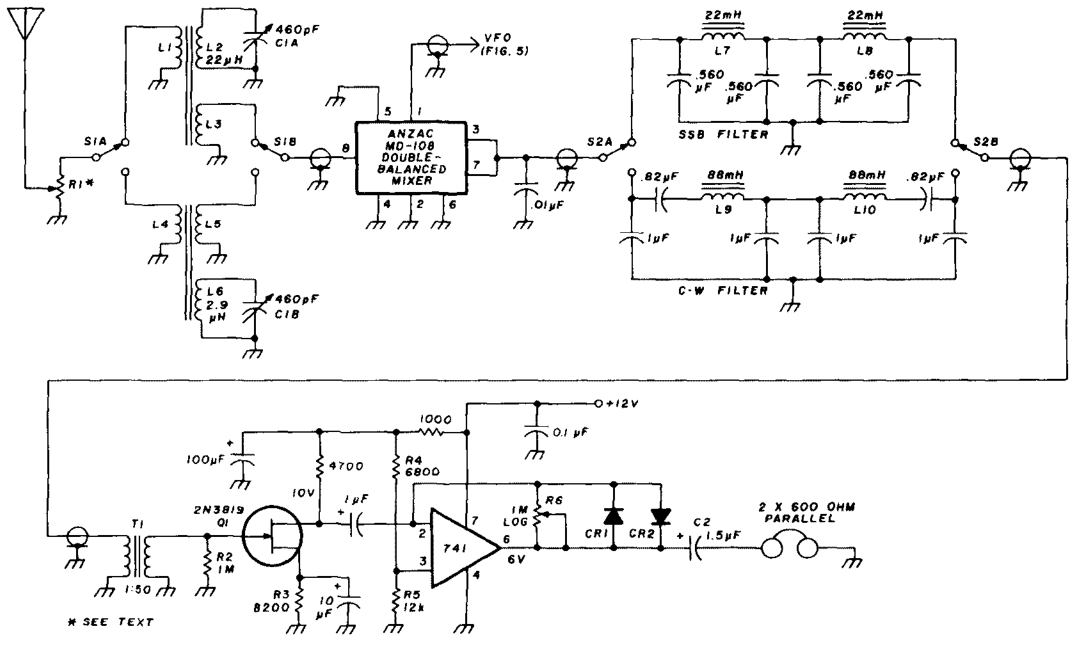



**fig. 3. Schematic of the direct-conversion receiver for 1.8-21 MHz less the vfo. Details of L1-L6 are described in the text. L7, L8, 88-mH toroids with windings series connected. S1, S2, dpdt toggle switch of good quality. T1, microphone transformer from tube-type transmitter. R3, selected for a dc drain voltage on Q1 of 10 volts; see text. The MD108 double-balanced mixer is made by Anzac. Pin numbers on the 741 op amp relate to the round (TO-5) package.**

**table 1. Component values for the vfo.**

| Band | Range (MHz) | L1 (μH) | C1 (pF) | C2 (pF) | C3 (pF) | C4 (pF) | C5 (pF) | C6 (pF) | R1 (ohms) | R2 (ohms) |
|---|---|---|---|---|---|---|---|---|---|---|
| 160 | 1.8-2.0 | 25 | 70 max | 39 | 390 | 800 | 800 | 390 | 0 | 220 |
| 80 | 3.5-3.8 | 5.7 | 70 max | 100 | 390 | 800 | 800 | 330 | 0 | 150 |
| 40 | 7.0-7.1 | 2.7 | 9 max | 0 | 300 | 800 | 800 | 160 | 120 | 150 |
| 20 | 14.0-14.35 | 1.7 | 9 max | 0 | 100 | 220 | 330 | 82 | 0 | 150 |
| 15 | 21.0-21.45 | 0.58 | 9 max | 0 | 220 | 220 | 330 | 47 | 0 | 150 |

selectivity will be poor. A few trials will quickly bring a good compromise between these conflicting requirements.

## variable-frequency oscillator

Oscillator frequency determines the tuning range of the receiver, so frequency stability is perhaps the most important oscillator requirement. The oscillator signal should have minimum noise modulation. As explained earlier, oscillator noise can become the limiting factor of receiver sensitivity. In my limited experience, an fet produces less noise than a bipolar transistor. Output power is of no consideration in a vfo; this is the job of the following amplifier, which is covered in a later part of the article.

For a multiband, direct-conversion receiver the oscillator signal should be generated directly, without premixing or frequency multiplication. This reduces the risk of spurious signals and noise modulation. It is, of couse, not a simple task to limit frequency drift on the higher-frequency bands, but

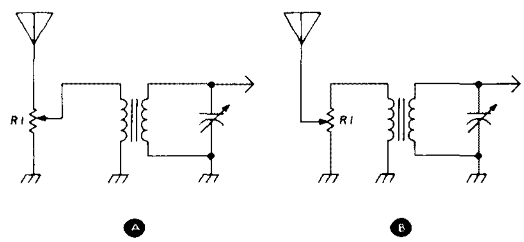



**fig. 4. Two ways of connecting a so-called "antenna attenuator." Method in (B) is preferred.**

with care it can be accomplished to a satisfactory degree. An example is the Atlas transceiver.

The type of circuit chosen for the oscillator is less important than the way it's implemented. Many good designs for vfos are found in the amateur literature. My solution to the vfo problem is offered only as an illustration. It contains parts that were available in my junkbox and should *not* be considered as the one and only acceptable circuit.

The vfo schematic is shown on the left in **fig. 5**. It includes five separate vfos, one for each band, 1.8 through 21 MHz. Each oscillator is tuned by a section of C1, a five-gang variable capacitor of 70 pF per section. For the 40-, 20-, and 15-meter oscillators, C1 was modified to a single plate in the rotor. (The 40-meter band in the Netherlands comprises only 7000-7100 kHz; the 80-meter band 3500-3800 kHz).

**Table 1** lists the values of L1, capacitors C1-C6, and R1, R2. (More about C6 later). Even if you have exactly the same values for C1-C5 and L1 available, there's a good chance that one or more of your oscillators won't oscillate. In this case your fets probably have less transconductance than the ones I happened to have. It's therefore better not to follow my design blindly but to use the parts you have on hand or can obtain, then tailor the circuit to your own requirements. A few guidelines: In principle C4 and C5 should be as large as possible and C3 as small as possible. This will provide loose coupling between fet and tuned circuit and will minimize the influence of the fet on the oscillation frequency. If you go too far in this direction, feedback will become too small and oscillation will stop. For a start, C4, C5 can be made equal.

The inductance of L1 is determined by the lowest-frequency in the band desired. L1 should resonate on that frequency with the total capacitance in parallel (C1 set to maximum). Variation of C1 with respect to the total fixed parallel capacitance determines the bandspread. This means fiddling with the capacitors and coil until the proper frequency range is covered while the oscillator still oscillates.

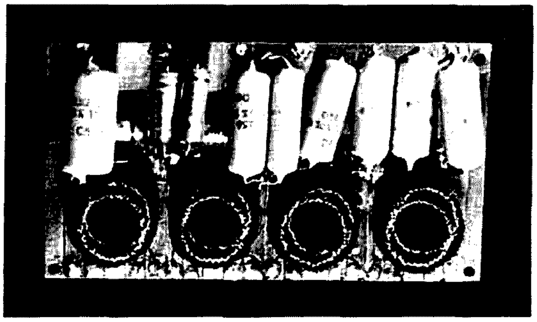
It seems that many amateurs shy away from this approach. But in practice it's not as bad as it sounds so long as you're well aware of what you're doing. Typical of this is the fact that the values of L1 in **table 1** were not calculated in the design stage, nor were they measured. I computed them afterward for this article from the frequency ranges covered and the capacitor values, so these values are of very limited accuracy.

Supply voltage should be as low as possible consistent with reliable oscillator starting. I found 5 volts to be a good compromise. Current increases quickly with voltage; and as dissipation in the fet increases with voltage *and* current, frequency stability suffers. An improvement suggested by PA0TW and PA0HWE, which I haven't tried, is to use a small preset pot between the cold end of the rf choke in the source lead and ground (**fig. 5**). This seems to give smooth oscillation control. It is absolutely necessary to stabilize the supply voltage by a zener,

with an electrolytic capacitor in parallel to suppress noise from the zener.

Output is taken from the drain through a small resistor. This method is due to DJ1BZ and features minimum effect on oscillation frequency. Because the fet operates in class C, current pulses flow through R2, and the resulting voltage is far from sinusoidal. We can improve the waveshape — again according to DJ1BZ — by including C6. It is given such a value that its reactance is roughly equal to the value of R2 in ohms for the highest frequency of the band concerned. With the component values as in **fig. 5** oscillator output voltage is about 0.5 V rms.

R1 is required only in case of vhf parasitics, which manifest themselves in erratic frequency jumps when rotating C1. R1 should not be any higher than necessary for good suppression of parasitics. In my



**Ssb and CW filter are combined on one piece of epoxy PC board. Toroids were first individually cemented to small pieces of board with epoxy. Copper plating was divided into five wide tracks with a knife. Winding ends were soldered to the ends of the tracks (center one not used), and the external connections were then made to the other ends. Small boards carrying the coils were then epoxied to the mother board.**

case R1 was needed only in the 40-meter vfo, but this requirement is unpredictable.

The oscillator required for a band is activated by connecting the supply volttage to that oscillator. Diodes CR1 function as OR gates. Only CR1 of the selected vfo conducts and connects the rf output to the buffer amplifier. The other diodes are reverse biased. Any point-contact silicon diode is suitable. (I used unidentified ones salvaged from a computer board).

It is necessary to shield the vfos well. This is even more important when the DC receiver forms part of a QRP CW transceiver. The smallest leak of transmitter output into the vfo will cause frequency pulling and/or a bad note. Watch the shaft of C1: it can easily act as an antenna and allow rf into the shield box. The remedy is to use an insulated shaft or shaft coupler.

I made a box from aluminum sheet and rectangular stock. Although no admirer of printed circuits for home construction, I put the five vfos on a piece of epoxy board to save room. The photographs show packaging details. A good slow-motion drive on C1 is recommended. The one I used came from one of the popular (at least in Europe) war surplus SCR-193 tuning units. It has a 1:50 worm drive. Originally the worm was driven by a thumbwheel protruding through the front panel. Because this method of tuning is very tiring when done regularly, I moved the unit through 90 degrees. The shaft of the worm was extended so it can protrude through the side wall of the vfo box and the front panel of the receiver where it carries a big knob with a crank. The extension shaft was cemented to the worm shaft with epoxy.

The slow-motion drive fits one end of the capacitor shaft. The other end of the shaft protrudes through the end wall of the shield box. A drum-type frequency dial can be fitted to this shaft end with a separate frequency calibration for each of the bands. The dial can be *read* through a window in the receiver front panel.

## buffer amplifier

The buffer amplifier increases the power level of the oscillator signal to about 5 milliwatts, which is needed by the MD108 double-balanced mixer. The circuit diagram is shown in the right-hand part of **fig. 5**. It is a broadband amplifier with two stages. The first stage with Q2 has series-negative feedback by nondecoupled emitter resistor R3. This causes both high input and output impedance, so Q2 causes negligible oscillator loading. Moreover its high input impedance is in parallel with the relatively low-valued resistors R2. The second stage with Q3 has shunt-negative feedback through R4, which also acts as collector resistor for Q2. Q3 therefore has low input and output impedance. The high output impedance of Q2 working into the low input impedance of Q3 causes a high mismatch, but it has the advantage that the two stages can be designed independently. Because of the low output impedance of Q3, variations in the loading impedance hardly affects amplifier operation.

Voltage amplification of the circuit is almost completely set by the ratio of R4 to R3 and is, to a large degree, independent of transistor characteristics, frequency, and supply voltage. This method of making broadband amplifiers with stages having alternate series- and shunt-negative feedback is due to E.M. Cherry and D.E. Hooper.[2] The simple approach by Cherry and Hooper allows the amateur to design good wideband amplifiers without too much computation and/or test gear.

T1 matches the output of Q3 to the LO input port of the double-balanced mixer. This is an aspect that

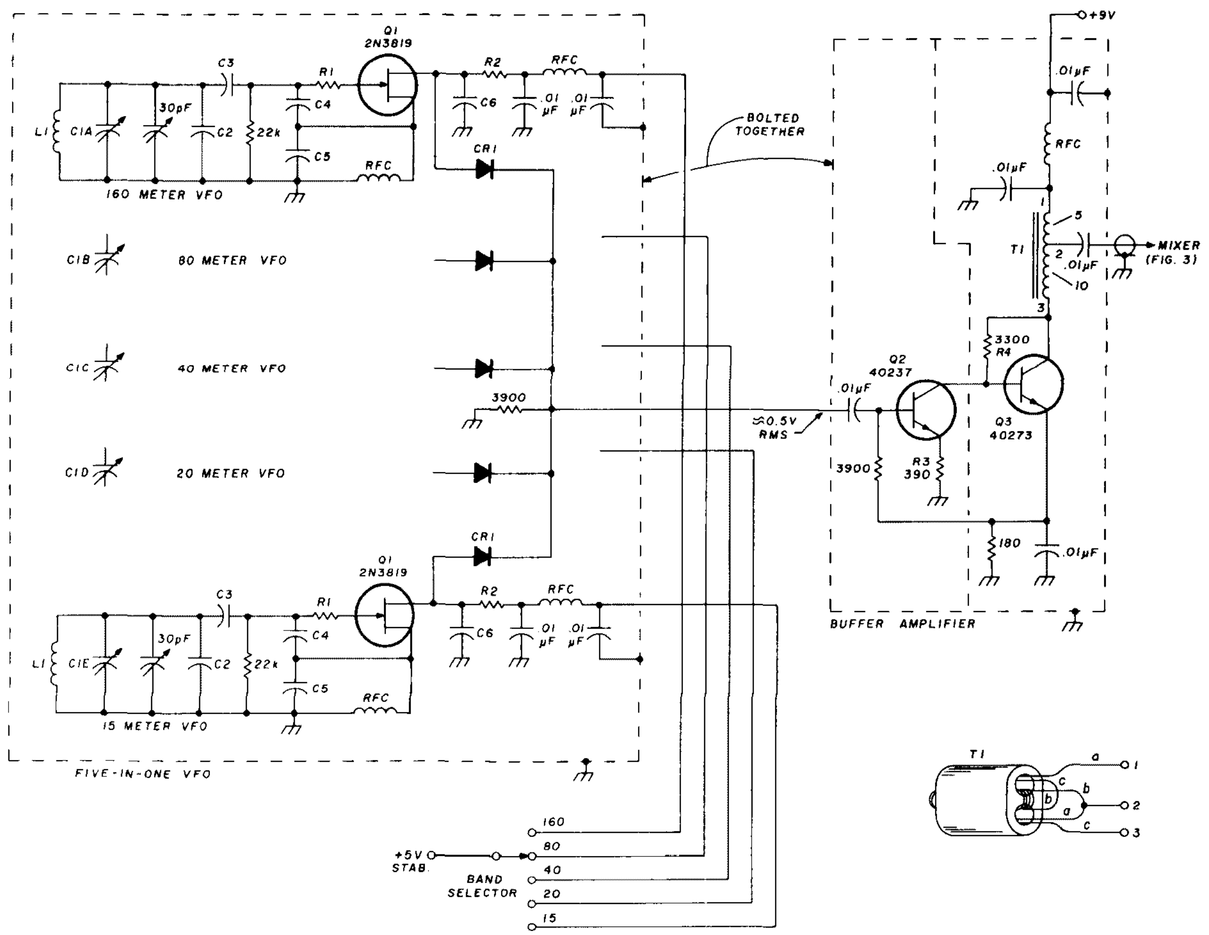



**fig. 5. Vfos and buffer amplifier schematic. A separate oscillator is used for each band, which is selected by connection to the supply voltage. C1 is a five-gang capacitor with all but one plate removed in the rotor section for 7, 14, and 21 MHz. CR1 can be any rf silicon diode such as a 1N914. Instead of the outdated RCA transistors, any modern rf silicon device can be used for Q2, Q3. Rf chokes are not critical; any type of 70 µH or higher inductance will do. Construction of wideband transformer T1 is shown at bottom; see text also.**

does not get the attention it deserves in many published designs for DC receivers. For proper mixer operation it is important that LO drive be sufficiently strong; some overdrive is less harmful than insufficient drive. For the MD108, the manufacturer specifies mixer characteristics at a LO drive of 7 dBm; that is, 5 milliwatts. Input impedance of the LO port is 50 ohms. Before we can decide on the stepdown ratio of T1, we need to know the optimum load impedance of Q3. The ac voltage at Q3's collector is set by the oscillator output voltage and the buffer amplifier voltage amplification. The unloaded voltage at the collector turns out to be about 2.6V.

Maximum power is delivered to the mixer when the *ac* through Q3 is maximum. The direct current is 10 mA. The rms current at the collector can therefore be about 7 mA at maximum (the peak value is then 10 mA). The optimum load impedance is that which passes 7 mA at 2.6V, which works out to be 371 ohms. So the stepdown voltage ratio of T1 should be $\sqrt{371/50} = 2.72$. Because of the construction of T1, the ratio can only be a whole number so we choose 3 as the nearest.

Construction of T1 is shown in **fig. 5**. T1 is an autotransformer with a trifilar winding. The core is powdered iron with two holes, as used in Europe for balun transformers in the input circuit of TV receivers. No doubt a suitable toroidal core of powdered iron or ferrite would do just as well. The number of turns is governed by the requirement of sufficient inductance at the lowest frequency to be used. Somewhat arbitrarily I decided that the inductive reactance of the winding between connections 1 and 2 should be four times the load impedance: 50 ohms at 1.8 MHz. A sample winding on the core in parallel with a known capacitor and coupled to a grid dipper revealed that five turns would be required between connection 1 and 2. To

obtain a transformer ratio of 3 a total of 15 turns will be necessary, of which 10 are between connections 2 and 3. I used silk-covered enamel wire of 32 AWG (0.2mm) because it happened to be available. Three pieces of wire are twisted, which is done conveniently with a hand drill, and the "rope" so formed is put five times through one hole of the core and back through the other. Connections are then made as indicated in **fig. 5**. Make the connections as *near to the* core as possible. The completed transformer is epoxy cemented to the buffer-amplifier shield box.

Any small box can be used for the buffer amplifier. The box is screwed to the oscillator housing, as shown in one of the photographs. A small feedthrough carries the signal from the oscillators into the buffer-amplifier box. As an extra precaution against feedback I put a partition between the two stages of the amplifier, as indicated in **fig. 5**. However, as the voltage amplification between Q2 base and Q3 collector is only some five times, this extra shielding is perhaps unnecessary. It is recommended to feed the buffer amplifier from a zener-stabilized 9-volt supply.

It is very important that the buffer stages operate within the linear part of their transfer characteristic. As soon as a transistor bottoms out the isolation offered is gone. This can be easily checked: putting a load on the amplifier in fig. 5 should have little or no effect on frequency. If frequency shift occurs and shielding and decoupling are alright, then overdrive of the buffer may be the cause. Decreasing R2 lowers the input voltage to the buffer. If this appears to be the remedy then C6 has to be corrected as well, as explained earlier.

## audio filters

I began my experiments with DC receivers usng an SL640 IC mixer made by Plessey, which has 350 ohms output impedance. The mixer was followed by a filter for ssb with a 2700-Hz cutoff frequency. The filter was a so-called Cauer or elliptic function design, which offers the steepest possible transition between passband and stopband with a given number of coils and capacitors. Its disadvantage is that all components have odd values, so coils must be tailor-made by paralleling pot cores and capacitors of standard values. The filter did an excellent job, however.

I replaced the SL640 with a double-balanced mixer using OA154Q germanium diodes, which had a measured output impedance of about 125 ohms. Since the filter had to be redesigned for the different impedance, I decided to use the well-known 88-mH toroids for coils with standard-value capacitors. This would make duplication by others easier. The filter was accordingly designed to the rules of classical image-parameter filter theory.

These filters do not have the steep transition between passband and stopband offered by modern filters, so some compensation was sought by lowering the cutoff frequency to about 2000 Hz, which is sufficient for ssb reception. After some trials, the filter of **fig. 6** emerged. The 22-mH coils were made from 88-mH toroids by placing the two windings

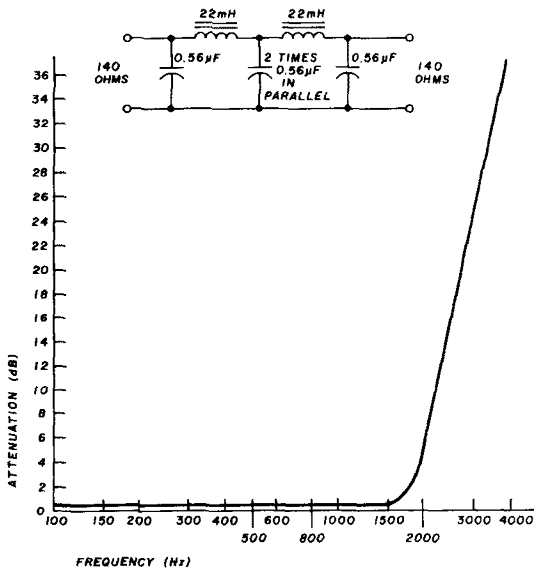



**fig. 6. Measured frequency response of ssb filter.**

in parallel.* It is important that the coils are connected so that the two windings don't oppose each other. In the filter passband attenuation is of the order of 0.5 dB; *cutoff* occurs near 1900 Hz. If you wish to check the design, the filter prototype can be found in reference **4**. It is built from four constant-k half sections.

For the CW filter I concluded that the passband should not be too narrow, because not only is it difficult to get the signal within the passband and keep it there but the tone is always of the same pitch, which becomes tiring during prolonged listening. This fact led to the filter of **fig. 7**, which consists of four 3-element series half sections. The passband ( − 3 dB) is between about 580 and 900 Hz. In musical terms, this means that the signal can be tuned through about a fifth without noticeable variation in amplitude. Attenuation increases faster on the high-frequency side of the passband. The CW filter is also useful as an outboard unit between any receiver and headphones in case the receiver or transceiver has insufficient i-f selectivity for CW use. Some resist-

*See also reference 3, which provides curves showing inductance values that can be obtained by removing turns from these popular surplus inductors.

ance padding will probably be required to match the filter to receiver output and headphone impedance. The signal loss that this entails is usually not serious, as in most cases more than enough signal is available for headphone operation. A breakdown of the CW filter in its four half dections is shown in **fig. 8**. The four coils, L1, combine to two coils of inductance 2L1.

By way of contrast, **fig. 9** shows a bandpass filter as used in some DC receivers published in *QST* and the ARRL Handbook. With two coils only *two* half sections are realized. The slope of the attenuation curve outside the passband is therefore only half that obtained by the filter of **fig. 8**. Another disadvantage of the filter in **fig. 9** is that the steepest slope is on the low-frequency side of the passband. Indeed, the same filter characteristic could have been obtained with only *one* coil if the other ends of the half sections in **fig. 9** had been joined. The coil in that case would have become ½ L2 in value.

The source and termination resistance of the filters should be 140 ohms for the ssb filter and 104 ohms for the CW filter. This reasonably matched the 125-balanced mixer. Later I changed to the Anzac MD108

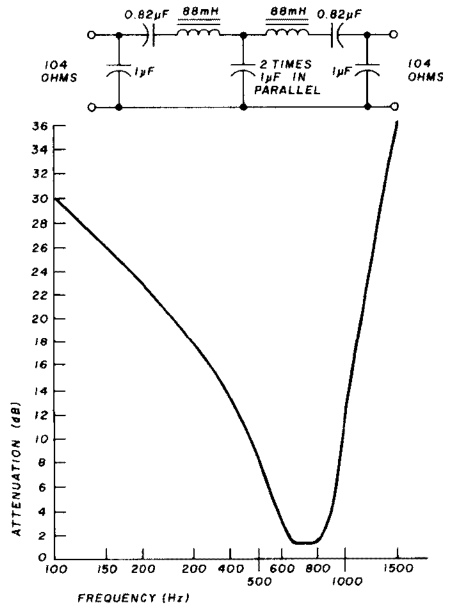



**fig. 7. Measured frequency response of CW filter.**

which has only 50 ohms output impedance. Luckily, filters based on the image parameter design method are a compromise on matching anyway, and in practice the mismatch on the input side does not noticeably detract from filter performance. Output matching is covered in the next paragraph.

Sometimes I'm asked why I use old fashioned coil-capacitor filters now that active filters without coils are available. In the first place I'm not sure that active filters with performance equal to my passive filters would be so simple; I'm afraid a considerable number of components would be needed. But I have also a more fundamental objection to the use of active filters in this particular application. The spectrum of signals offered to the filter by the mixer comprises a large dynamic range that could easily be some 80 dB or more. I fear this range is more than an active filter can handle. Either the weakest signals will drown in the noise of the device or the strongest will overload it. One should not forget that, especially in the case of steep cutoff filters, some parts of the circuit will carry much higher voltages than appear on input and output terminals — and at those points the danger of overloading is greatest. Therefore I prefer the classic LC filter. Construction should prove no problem at and the 88-mH toroids are inexpensive and plentiful.

## audio-frequency amplifier

In a DC receiver the input signal is only *attenuated* in the first stages. Amplification occurs for the first time after the af filters, so signal power reaches a minimum at the input of the af amplifier. Unless the mixer is poor and oscillator noise dominates, the receiver signal-to-noise ratio is determined by the af amplifier input stage. I again get the impression that this important consideration did not always receive the attention it deserved in some of the DC-receiver designs I've seen.

If a bipolar transistor is used in the first stage of the af amplifier, it should be a low-noise device; e.g., a type suitable for the input stage of a tape or cassette recorder. From the manufacturer's data sheet one can find the collector current for optimum noise factor, usually some tens of microamps. But these sheets show another fact: optimum noise factor is obtained with a specific output resistance of the signal source for feeding the transistor! Agreed, the curve for noise factor as a function of source resistance shows a rather broad minimum, but if the af filter is connected directly to the output of the af filters, as is often done, noise mismatch may be so serious that s/n ratio is degraded by several dB.

Professional designers of low-level af amplifiers use input transformers to obtain an optimum noise match if the source resistance differs widely from the optimum, as in the case of a dynamic microphone for example. We could do the same in our DC receiver. Source resistance in this case is the 50-ohm output resistance of the double-balanced mixer as seen through the af filters over the major part of the filter

passband. Optimum source resistance for the af input transistor depends on device type and collector current and can vary between a few thousand and several tens of thousand ohms.

From these data a suitable transformer can be specified. But would you have one, or could you buy it somewhere? Perhaps. But don't run out to get one because there is another snag. Not only should the af amplifier have a certain resistance at its input; at the same time, the af filter in use should have the proper termination resistance at its output. The load on the filter will be the input resistance of the af amplifier, transformed by the transformer — so this requirement also fixes the transformer ratio. It would be most unlikely that the transformer ratio so found would coincide with the one for optimum noise matching, so we have a problem.

It would be nice if we could choose a transformer with an audio generator and af voltmeter showed the voltage step up from primary to secondary to be of the order of fifty. So signal-to-noise ratio is raised fifty times compared to a straight connection without a transformer.

The matter of filter termination is still to be settled. This is simply accounted for by R2 in parallel with the secondary winding. R2 is transformed to the input side of the transformer, divided by the square of the transformer ratio. The value of 1 megohm was suitable in my case. The filters are somewhat underloaded at their outputs, but signal from the filters is larger than with proper termination. As stated before, filter matching is not very critical.

The remainder of the af amplifier is simple. There is considerable spread in the characteristics of fets. It is therefore better to use a variable resistor as source resistor R3 first and to adjust it for a dc voltage of 10

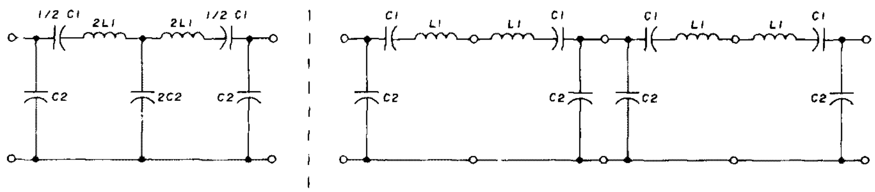



**fig. 8. The CW filter is a combination of four half sections, shown to the right of the vertical line. Note how coils and capacitors in the individual sections are combined in the actual filter at the left of the dashed vertical line.**

ratio for best s/n ratio at the af-input stage and control filter termination separately. We can by using an fet. An even better s/n ratio than with a bipolar transistor also occurs. The input resistance of an fet at audio frequencies can be considered infinite for our purpose, so there's no loading at the input. Noise of an fet is lower than in even low-noise bipolar transistors. Noise in the fet can be thought of as being generated in a noise-voltage source in series with the input (gate). The higher the signal input voltage is made, the better the s/n ratio becomes, as signal and noise voltage operate in series on the gate. So the higher the step up of the transformer between af filter and af amplifier, the better the s/n ratio becomes. The limiting factor is the transformer itself. With high ratios the capacitance of the secondary winding limits the high-frequency response. But this is of more concern to the hi-fi equipment designer, as the highest audio frequency we're interested in is about 2000 Hz.

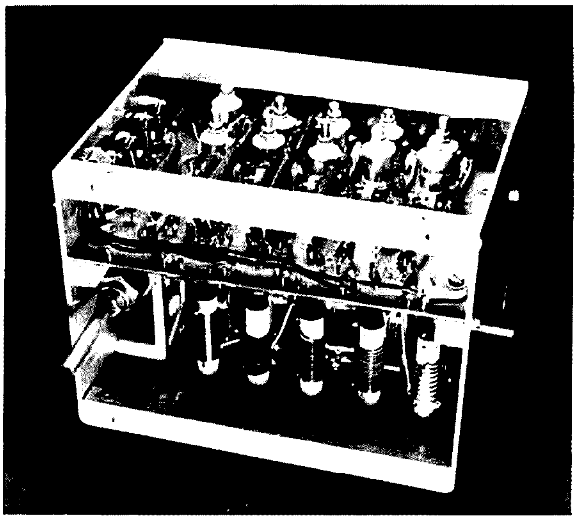
Very suitable for our purpose are microphone transformers from communications equipment using tubes (war-surplus transmitters, obsolete mobile radios). Transformer T1 in **fig. 3** came from Wireless Set type 19, a famous British WW II veteran. The photograph shows the unit in its shielded box. A test volts at the drain of Q1. The resistor is then measured and a fixed resistor of nearest standard value substituted.

The major share of the total amplification is provided by the popular 741 op amp. Resistors R4 and R5 are selected so that dc voltage at the output (pin 6) is half the supply voltage; i.e., 6 volts. If necessary R4 and/or R5 can be changed if the voltage at pin 6 differs too much from half supply voltage. Volume control R6 changes the feedback; this somewhat unusual system has advantages with op amps. Silicon diodes CR1 and CR2 will protect your ears in case a strong signal appears unexpectedly; they limit output voltage to a maximum of about 1.2 volt peak to peak.

Because the output resistance of an op amp is so low and becomes even less with feedback, connecting the diodes in parallel with the headphones would not be very effective. That's why they have been incorporated in the feedback circuit. When 600-ohm stereo headphones are used with both halves in parallel, the diodes don't conduct at normal listening level. If you prefer pop-group sound level, better omit CR1 and CR2.

When starting my tests with direct conversion I used a speaker. Instead of the 741 op amp I tried a

Siemens TAA300 and the Plessey SL630 as output IC amplifiers, but I like the sound from good headphones much better. Modern stereo headphones have very good bass reproduction, even very light hum is reproduced faithfully. It's difficult to avoid hum induction completely, because T1 is sensitive to the stray magnetic fields of power-line transformers, even if a foot (30cm) away. For this reason C2, in series with the output, is made rather small. The frequency response falls at 6 dB per octave below 350 Hz. This is also useful at ssb as the top is cut at 2000 Hz; attenuating the lows as well restores the tone balance.



**Inside view of the five-in-one vfo.**

## agc

In the final version of my receiver, shown in **fig. 3**, no agc is used. Whether or not agc is desirable depends mainly on the use of the receiver. For CW it's not necessary. Only the more sophisticated forms, like hang agc, contribute to operating convenience. On ssb agc is not necessary either but is nice to have.

I have tried agc in my DC receiver. For this purpose the mixer was preceded by a Plessey SL610 rf amplifier. This IC can be directly connected to the mixer input without using a tuned circuit or other matching device. Audio agc voltage was generated by another Plessey IC, the SL621. This is a sophisticated circuit, providing hang agc without "hang" on short noise or interference bursts.

The SL610 has a control range of 50 dB. Of course this is not enough for a full-fledged agc, but it was sufficient for the range of signals appearing within one frequency band at a certain time. Control was very pleasant, but a nasty side effect spoiled performance: when tuning a strong carrier the agc voltage increased in a series of steps. I tried changing the time constants in the loop but nothing helped. Also the Plessey application engineer could not think of a remedy. It is clear that the SL610-SL621 combination was designed for the superhet — in which they perform excellently — but in the DC receiver they obviously do not feel at home. Eventually I dropped agc. First, the rf amplifier was not needed from a sensitivity point of view, and secondly my receiver is part of a QRP telegraphy transceiver and is very seldom used for listening to ssb.

## tunable hum

Many DC receivers have hum that appears when the input circuit is tuned to the signal (oscillator) frequency. I think the explanation is that some oscillator power leaks through the mixer and finds its way to the input circuit. The voltage on the circuit peaks when the circuit is brought to resonance. Some of the power is fed into the antenna and radiated. If the antenna is unsymmetrical and works against ground, the antenna current also flows over the ground connection. Some or all of this path to ground is through power supply and ac line. The current finds its way to the ac line through the rectifier diodes in the supply and the capacitance between primary and secondary windings of the power transformer. But the diodes operate as switches that open and close 60 times a second (50 times in Europe), so the rf current is chopped (modulated) at this frequency as well. It is clear that because of this we find an rf signal on the input circuit that is amplitude modulated at line frequency. This is treated like any "normal" signal and finds its way through the receiver; the hum is demodulated in the mixer and becomes audible at the output.

The remedies are clear. A good solution is to use a 12-volt battery for primary power. With a total consumption of 25 mA, this is an attractive alternative. Of course a good ground connection is a must on battery-operated gear. Another possibility is to use an antenna that does not depend on a ground connection for its operation. When using my inverted vee with open-line feeders and an antenna tuner there is no trace of tuned hum. Still another solution is to use a good ground separate from the ac line. This alone is not sufficient; the path through the power supply should be made more unattractive to rf by putting an rf choke between receiver and power supply.

## input-circuit microphonics

This effect is related to the previous one. Some

oscillator voltage is present on the input tuned circuit. If variable capacitor C1 is not mechanically rigid, the plates may vibrate. The resulting capacitance variations modulate the rf signal in the circuit both in amplitude and phase. The signal is demodulated in the mixer and the sound emanates from headphones or speaker. If the speaker is near C1, an acoustic howl may be set up in extreme cases. Apart from

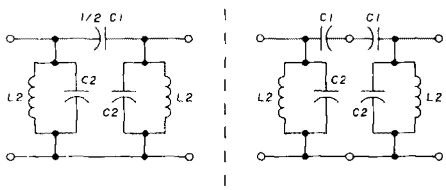



**fig. 9. Bandpass filter built from two half sections provides only half as steep a filter slope compared with circuit of fig. 8, although the number of coils is two in each case. The actual filter is to the left of the dotted line.**

using a good-quality capacitor for C1, it also helps to prevent the oscillato r signal from appearing on the input circuit. This depends on good balance in the mixer, shielding of oscillator and buffer circuit, and the connection to the mixer. With the Anzac MD108, microphonics were not a problem in my case.

## rf amplification, yes or no?

It is a real pleasure to operate the DC receiver. Signals stand out clearly against an almost quiet background. In poorly designed superheterodyne or DC receivers, even strong signals can sound blurred, almost always a sign that oscillator noise is modulating the incoming signals. Nothing of the sort happens in my receiver. Signals are as clear as a bell as soon as they are strong enough to override the noise.

The DC receiver seems much less "nervous" than a superhet, but it is only fair to admit that this is partly due to the absence of agc. Oldtimers, using a good DC receiver will no doubt be reminded of a tuned radio-frequency set, the old faithful of the twenties and early thirties, but without its peculiarities.

Sensitivity of my receiver, with an average 0.82 microvolt across 50 ohms for a 10 dB s+n/n ratio, may not seem impressive by modern standards and it is perhaps wise to give this matter a closer look. It would be nice to know the noise factor of the receiver. In my home lab I don't have a noise generator, so I can't measure noise factor directly, but it can be calculated from the measured sensitivity. It is necessary to know the noise bandwidth of the rig for this computation. This is an equivalent rectangular-shaped passband that, with the same height as the maximum of the real passband of the receiver, passes the same noise power. By taking the frequency response of the af portion of the receiver, the noise bandwidth can be determined from the response curve. The noise bandwidth so-found was 1340 Hz, using the ssb filter. The rf bandwidth is twice this value in a DC receiver, which accounts for the noise power that is passed.

Noise factor can now be determined by computation, or more easily, by the use of a suitable chart.[5] I obtained a noise factor of 15 dB at the mixer input. At the antenna terminals this will be a bit poorer because some signal is lost in the tuned input circuit. Assuming 2 dB for this loss, the noise factor at the receiver input is 17 dB. If this is not good enough, an rf amplifier is called for. But the strong-signal characteristics of the receiver — in particular a-m breakthrough, the most troublesome effect in a DC receiver — suffers just as many dB as the rf amplification you put in.

Is a 17-dB noise factor acceptable? To answer that question, refer to an excellent article by Jim Fisk, W1HR,[6] in which a table lists the maximum noise factor a receiver may have so that, in a quiet location, receiver noise is 3 dB less than noise from outside. Of course, external noise is dependent on many factors and varies widely with time , location, and frequency so W1HR rightly warns that no guarantee is given that receiver noise will *always* be 3 dB weaker than external noise.

The following figures are given for the hf-bands: 1.8 MHz, 45 dB; 3.5 MHz, 37 dB; 7 MHz, 27 dB; 14 MHz, 24 dB; 21 MHz, 20 dB; and 28 MHz, 15 dB. As my receiver does not cover 28 MHz a 17-dB noise factor can be considered acceptable. Practice at my reasonably quiet location seems to bear this out: even when 15 meters is dead a slight rise in noise from the antenna is noticed when the input circuit is tuned to the signal frequency. If the receiver covered 28 MHz as well, an rf amplifier would certainly be needed to catch the really weak ones.

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**ham radio**

# high performance 20-meter receiver with digital frequency readout

## Part two: digital counter, display, and receiver integration

**A digital display on modern receivers** is essential for accurate frequency determination of both received and transmitted signals. The division of the amateur bands into a multitude of operating modes, license restrictions, and accuracy in individual operator contacts makes the analog interpolating dial techniques difficult and leaves much room for frequency error. When crowded DX conditions prevail, the abililty to change frequency to a special receiving frequency is a must for working that "Once-in-a-decade" contact.

Numerous articles have appeared that use the variable high-frequency oscillator (HFO) as a base for determining the received-signal frequency. These systems use the HFO frequency and presettable decade counter schemes to display the received frequency in kHz. Most systems display either a 0-500 kHz or 500-999 kHz readout, depending on band segments, and the operator mentally supplies the significant Megahertz digits. An excellent method so far as it goes — however, since we're also changing our BFO frequency between USB, LSB and CW, and since small variations in the local-oscillator frequency of a dual-conversion receiver have thermal drifts, there is still considerable difficulty in determining *exactly* the received-signal frequency and the transmitting frequency when using the HFO as the only reference. Equally difficult is the accurate calibration of such a system, which is limited to the accuracy of only one oscillator reference, even after thermal equilibrium is achieved in the receiver system.

The digital display system described here provides accurate display of the received-signal frequency by counting the HFO, LO and BFO outputs, summing the counts from these oscillators, and displaying the actual received frequency. In addition, the basic BFO oscillator described in part one of this article[1] includes a nominal (455-kHz) oscillator mode for zero beating the received signal to determine the transmitted signal resting, or center, frequency and not necessarily the ssb or CW beat frequency.

The digital counter may be used with receivers that

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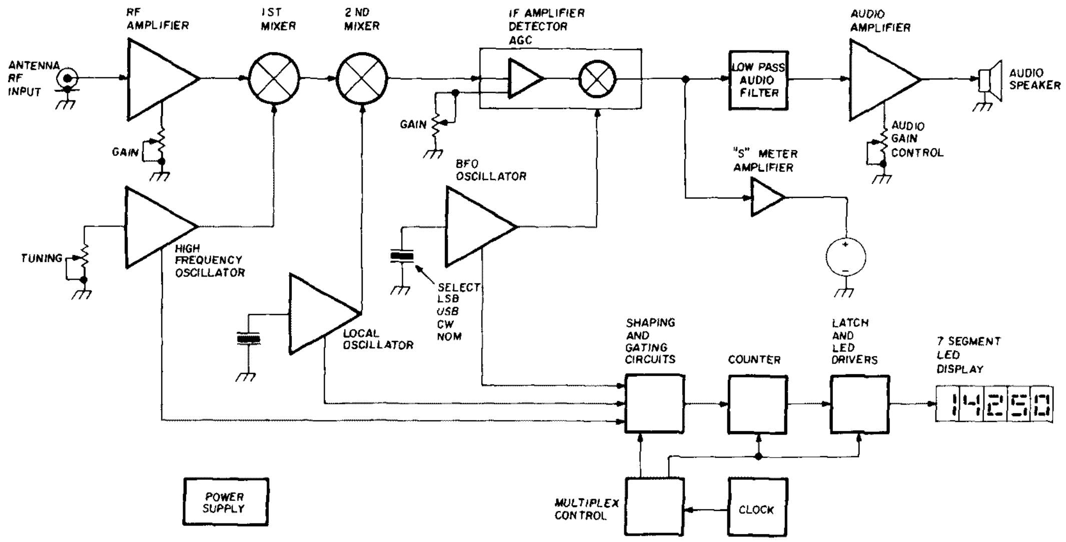



**fig. 1. Block diagram of 20-meter high-performance receiver.**

are not of the dual-conversion type by excluding the local-oscillator beat frequency input signal to the shaping circuits. The counter may also be used to monitor a transmitter frequency by simple diode switching, using only one of the buffer input circuits.

## the digital-display system

Consider that the signal being received is 14.125 MHz, as an example. From part one of this article, the first i-f is nominally 1500 kHz, and the second i-f is 455 kHz. We can sum the receiver oscillator outputs as follows:

| | |
|---|---|
| HFO | 12.625 MHz |
| LO | 1.045 MHz nominal |
| BFO | 0.455 MHz nominal |
| DISPLAY | 14.125 MHz |

The digital scheme that follows does exactly that — it counts each oscillator output individually, sums the counts, and displays the results.

As a reminder, **fig. 1** is a block diagram of the overall receiver. An input signal is obtained from each of the oscillator circuits, processed in a buffer gate circuit, then counted and displayed using a 5-digit, 7-segment LED readout. A clock circuit generates a stable reference for multiplex control, counting, and transfer of the updated frequency data at regular intervals.

The display update is 2.5 Hz. Several systems described in previous articles have used display rates of 10-20 Hz; however, my experience indicates that a fast display change rate is neither necessary nor desirable because of the flicker irritation. Display rate changes of 1-5 Hz seem psychologically less irritating and adequate for frequency readout. During rapid traversal of band-end to band-end, this error in frequency display is not objectionable, since familiarity with the counter operation and experience will quickly allow you to estimate the approximate stopping point between display rate changes for receiving a particular band segment.

**Counter**. The counter timing diagram is illustrated in **fig. 2**. For clarity, one full second of the system operation is shown. The basic clock frequency is 10 Hz. Each clock cycle is 100 milliseconds, and each display period is 400 milliseconds, at which time the display is updated and held for another 400-millisecond period. Considering an ideal 1-second case, during the first 100 milliseconds the HFO is counted; during the second 100 milliseconds the local oscillator is counted; and during the next 100 milliseconds the BFO is counted. The summed counts of the HFO, LO and BFO are then transferred to the display, and the counter is reset to start the cycle over again each 400 milliseconds. Transfer and display occur during the fourth 100-millisecond period of the control cycle.

The basic system is a ripple counter. To allow for all of the internal flip-flops of the counter to ripple through before transfer transition, a delay period is required before transfer of the ripple counter chain and reset. This action is accomplished by using a strobe pulse, which occurs during the multiplex $\overline{A}\,\overline{B}$ time period (**fig. 2**). The strobe pulse is derived by looking back into the 10-Hz clock decade counter and using the QC output, which occurs twice during

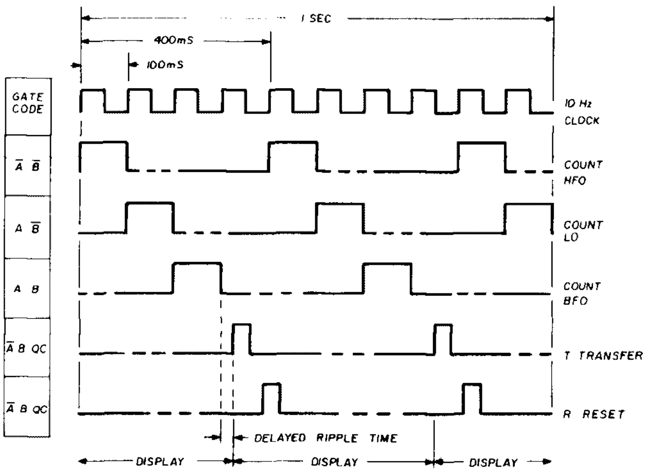



**fig. 2. Counter timing diagram for a 2.5 Hz count/display operation.**

the multiplex $\overline{A}$ B state. The strobe used in conjunction with the 10-Hz clock is then ganged to provide the transfer and reset pulses at the correct timing point.

The gate-code table in **fig. 2** illustrates the gating circuit codes that are generated from the multiplexer to accomplish each of the illustrated count-and-transfer activities. The gate codes are purely arbitrary and are not a part of device output terminal identification. They are merely a scheme for schematically illustrating the binary state of the controller and controller output.

Lower-power TTL and Schottky logic is used in this system. Selection of TTL and LS devices is presently the most cost-effective design in low part quantities such as this. A recent series of counter-latch LED driver chips has become available; however, for amateur procurement of low-quantity parts within reasonable delivery times, I chose to use a straightforward, easily duplicated design rather than one using hard-to-find ICs.

**Input buffer**. The oscillator outputs are buffered and shaped using the circuits shown in **fig. 3**. An fet buffer provides a high-input impedance to minimize loading, distortion, and frequency drift of previous oscillator stages. The fet-buffered output is capacitively coupled into a common-emitter driver, which is directly coupled to the input of a NAND gate shaper. The use of LS logic assures fast rise times and clean pulse trains from the oscillators. Generous use of decoupling capacitors and rf chokes minimizes feedback of spurious signals into the +12 volt source, which is common to the other rf and audio stages of the receiver.

**Gating system**. The gating technique is shown in **fig. 4**. The 3-input NAND gate always has one of the oscillator trains on its input. A multiplex generator logic level, tied into the other two inputs, either inhibits or allows the pulse train to pass through the gate. Assuming a pulse train appearing on the input of terminal C of the gate, both the A and B input lines must be in a logic one state (high) for the train to pass through the gate. If either the A or B input is low, the gate will be inhibited. It's important to understand that the NAND gate is an active low device. Each of the buffered and shaped oscillator lines is constantly looking into its individual 3-input NAND gate. The multiplex generator provides a series of logic levels in sequence to allow first one pulse train to pass, then the next, and finally the last oscillator train at the specified timing interval indicated in **fig. 2**.

The output of the multiplex-controlled gate of

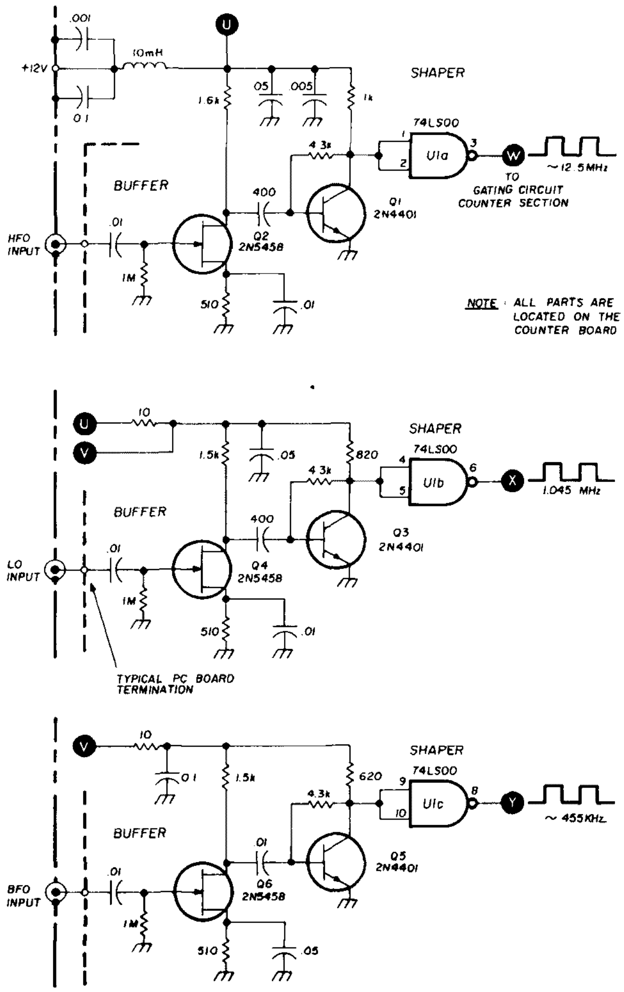



**fig. 3. Input buffer and shaping circuits, counter section.**

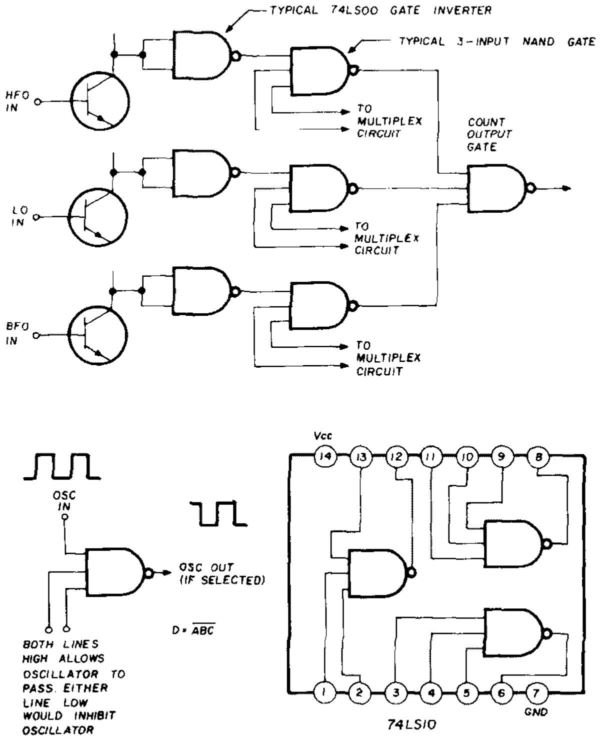



**fig. 4. Triple 3-input positive NAND gate function diagram.**

each oscillator train is, in turn, tied to the input of another 3-input NAND gate. Let's see how this works. Assume that the multiplex generator puts the A and B inputs of the HFO 3-input NAND gate high. The gate will allow the pulse train to proceed to the output and into the count-output gate. If the multiplexer is operating correctly, the A and/or B input of the LO and BFO 3-input NAND gate must have at least one terminal at logic zero to inhibit the train (don't confuse the A and B input terminals with the symbols A and B used in the controller binary state code). When the A and B input terminals are high, the input count gate will pass the HFO train. The next multiplex function will inhibit the HFO and BFO trains to pass the LO train of pulses, and subsequently the BFO pulse train. When the count gate is inhibited, the output C terminal will be high because of the active low characteristic of the device.

Because we're using active low outputs from the U2 gates (**fig. 5**), the 3-input NAND gate of U4 at pins 1, 2, and 13 will have at least two lines high during a count cycle *state, allowing the count to proceed to* U8. During the multiplexer state $\overline{A}\,\overline{B}$, the lines from U2 output are all high; thus pin 12 of U4, the gate output, is low and inhibits all pulse-train activity.

A complete schematic of the gating scheme is illustrated in **fig. 5**. The individual gates are controlled by the multiplexer to pass through a control gate and first decade counter. In addition, the multiplexer

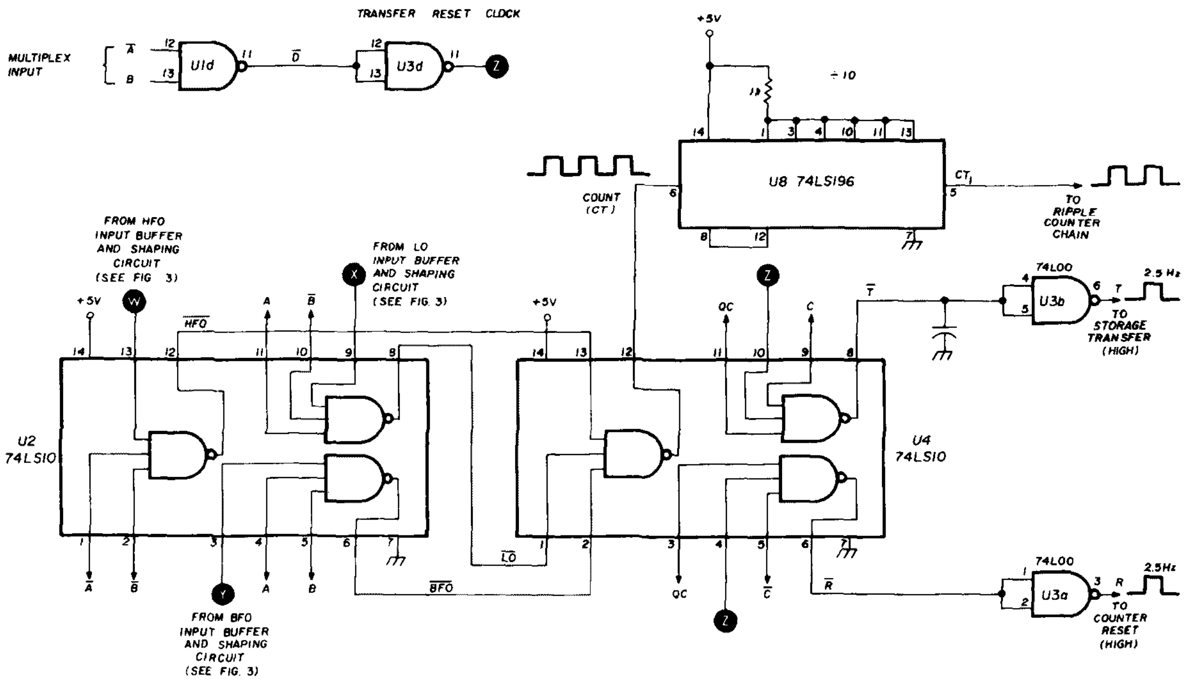



**fig. 5. Gating circuits; counter section schematic.**

generates a series of gating commands to two other 3-input NAND gates to generate the storage transfer and counter reset lines.

U1 and U3 gates provide the transfer reset clock, which is strobed by QC, clock (C), and NOT clock $(\overline{C})$, as shown in **fig. 5** using two 3-input NAND gates of U4. Since the output is active low, and transfer and reset occur on the leading edge of a pulse, two additional 2-input NAND gates (U3) invert the pulse to the correct polarity. The heart of the multiplexer system is a simple dual-D flip-flop (**fig. 6**), which generates a series of gate control pulses: 00, 01, 10, 11. These pulses, in combination with the clock, provide a continuous series of multiplex control lines: (000, 001, 010, . . . 111), so that the various counter states are derived for count, transfer, and reset at specific timing intervals.

**System clock generator**. A stable and accurate 10-Hz clock is generated from a 100-kHz crystal oscillator, which is then decade-divided to produce a basic clock-NOT clock pulse train, identified as C and $\overline{C}$. The circuit is illustrated in **fig. 7**. The entire accuracy of the system depends on the accuracy and stability of the clock. (This is *not* the place to use an inexpensive 100-kHz crystal!) Use extra effort in selecting a resonating capacitor, or use a small trimmer for frequency accuracy. Normal auto-ranging

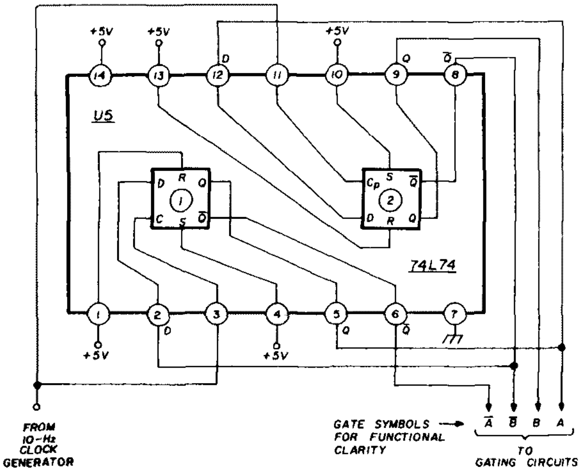



**fig. 6. Multiplex gate control generator; counter section schematic.**

calibrating the entire clock and receiver circuitry using WWV at 15-MHz is discussed in the final alignment and calibration section of this article.

**Ripple counter and display circuits**. The ripple counter, latch drivers, and display are shown in **fig. 8**. The system is straightforward, but let's see just what's happening. From **fig. 5** we have gated three

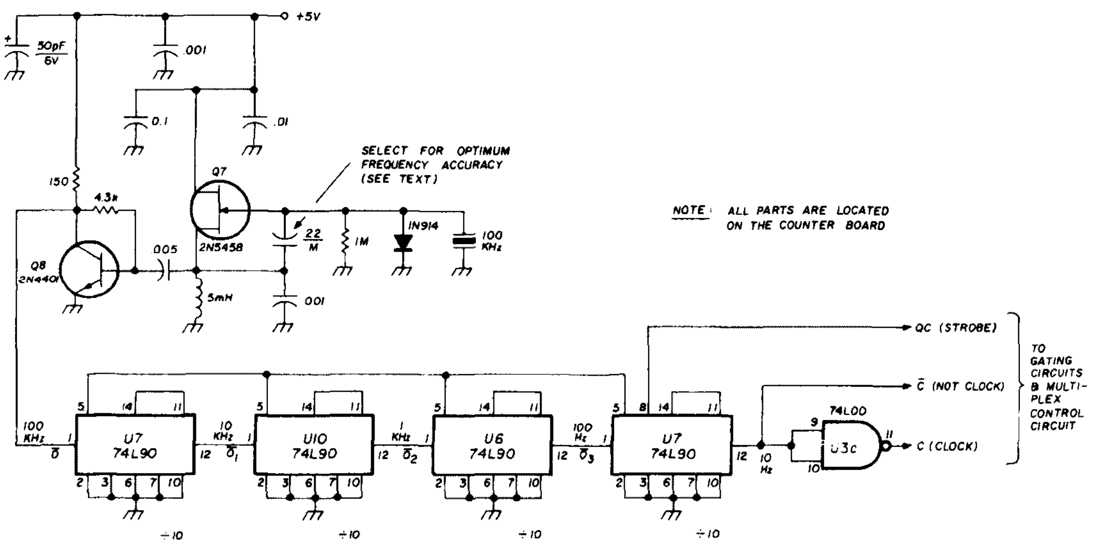



**fig. 7. 10-Hz clock and NOT-clock generator schematic.**

frequency counters will *not* provide sufficient accuracy for the clock oscillator frequency measurement. If possible, beg or borrow a frequency counter with 10- or 100-second gating capability and monitor the 10-Hz clock to ensure proper operation over extended time intervals. An alternative method of

input counts: the HFO, LO, and BFO. We sampled each of these pulse trains for 100 milliseconds. From our original example of the 14.125-MHz received signal, the HFO pulse train is 12.625 MHz divided by 100 milliseconds, or $1.2625 \times 10^6$ pulses; the LO and BFO are also sampled at the same period and are

thus $104.5 \times 10^3$ pulses respectively. The total accumulated counts after gating is the sum of the 100-millisecond-sampled oscillators, or:

$$\begin{array}{r} 1.2625 \times 10^6 \\ 0.1045 \times 10^6 \\ \underline{0.0455 \times 10^6} \\ 1.4125 \times 10^6 \end{array}$$

These are the counts passing through the count output gate to the first decade divider, U8 (**fig. 5**). After decade division, the ripple counter chain could be considered as viewing a count rate of 141.25 kHz in our example.

**Fig. 8** refers to the ripple counter input as $CT_1$. This 141.125 kHz train first passed through an additional decade divider to produce $CT_2$. In our example, this would now be a pulse train at a 14.125-kHz rate. **Fig. 9** shows the functional diagram of the ripple-counter action. $CT_1$ enters a divide-by-10 stage; its output is $CT_2$. $CT_2$ enters a decade divider

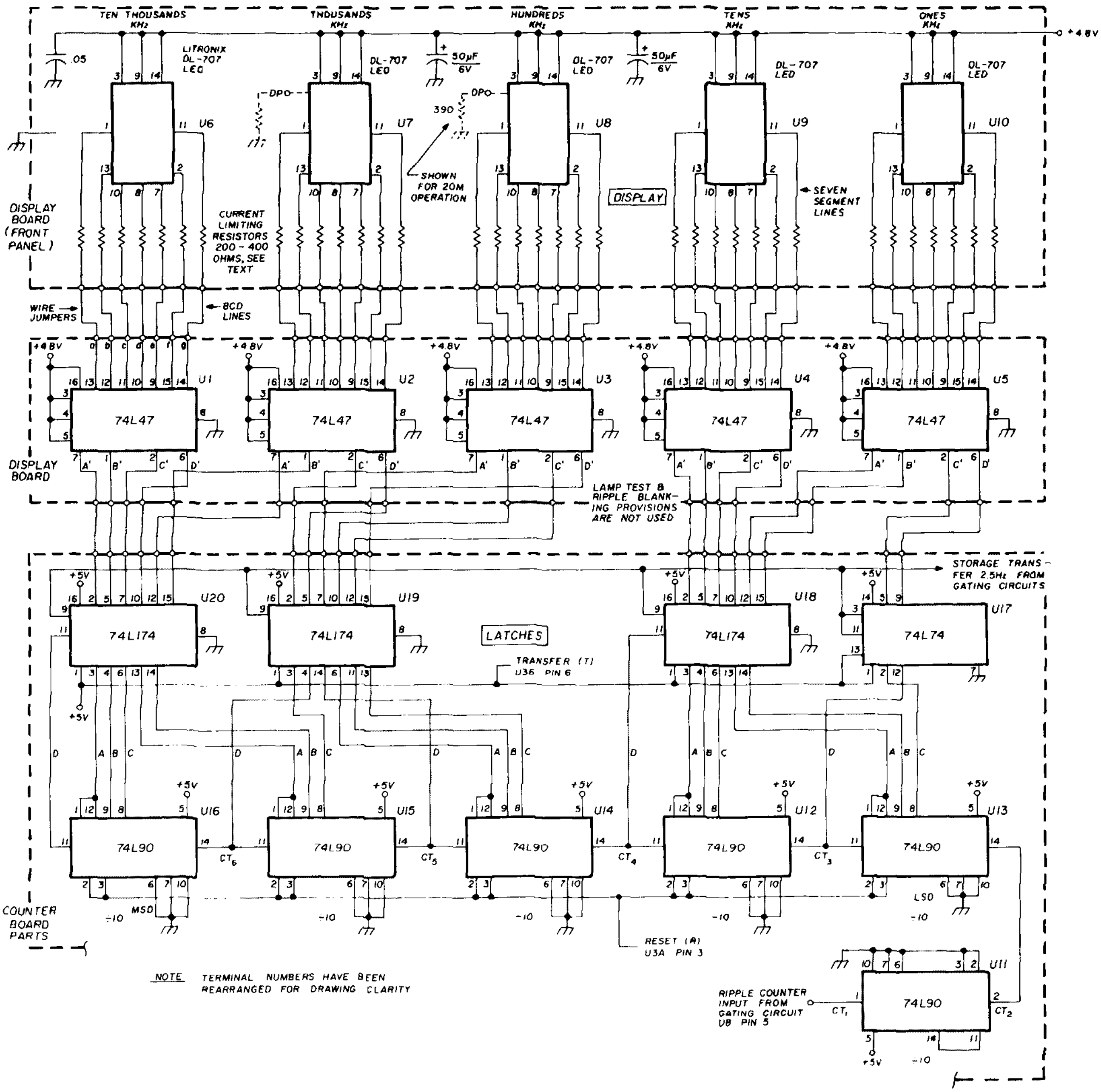



**fig. 8. Ripple counter, latch, and segment display schematic.**

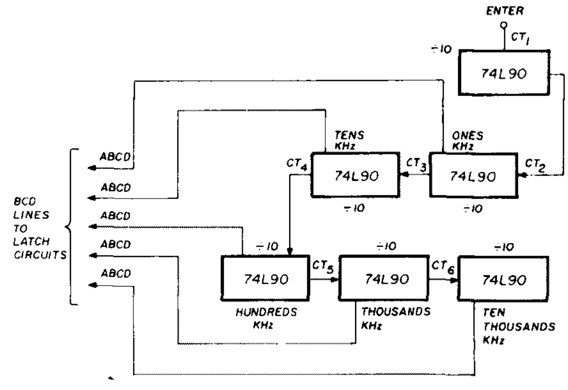



**fig. 9. Ripple counter functional diagram.**

with BCD ouput lines. Its output, $CT_3$, enters an additional decade divider with corresponding BCD output lines and so on until we reach the last or most-significant digit counter with its BCD lines.

**Fig. 10** is an abbreviated picture of the individual decade, latch, display driver, and display scheme. Each decade counter accumulates, in turn, the number of counts from the previous stages during the count cycle. Upon multiplex command, the storage register transfers the accumulated counts in each decade divider to the display driver; the storage-register output is held in this stage until the next transfer pulse. The display driver receives the BCD code from its corresponding 7490 counter and converts this binary code into a 7-segment LED format for numerical presentation.

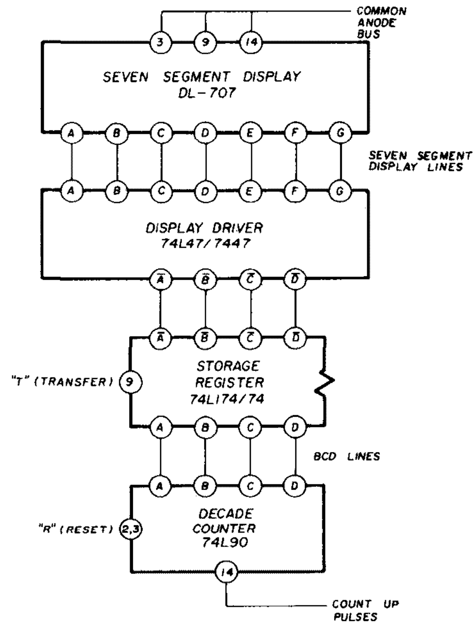
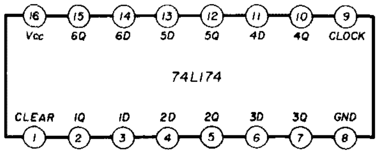



**fig. 10. Typical count and display scheme (single digit).**

Since the storage register is held in its state until the next transfer pulse, the counter (ripple) can be reset and will proceed to sample the oscillator chain again, as described previously. Upon arrival of the transfer pulse, this updated data is cycled again into the display. This count, transfer, display, and reset continually cycles each 400-millisecond period (or two and one-half times a second).

The latch circuit is the familiar D-type flip-flop with a clear input so that fresh data may enter the D input




INFORMATION ON THE D INPUT LINES IS TRANSFEREO TO THE Q OUTPUTS ON THE POSITIVE-GOING EDGE OF THE CLOCK PULSE. CLOCK FREQUENCY ~ 35MHz MAX.

| INPUTS | | | OUTPUTS |
|---|---|---|---|
| CLEAR | CLOCK | D | Q |
| L | X | X | H |
| H | ↑ | H | H |
| H | ↑ | L | L |
| H | L | X | $Q_0$ |

**fig. 11. HEX D-Type flip-flop with clear (top view).**

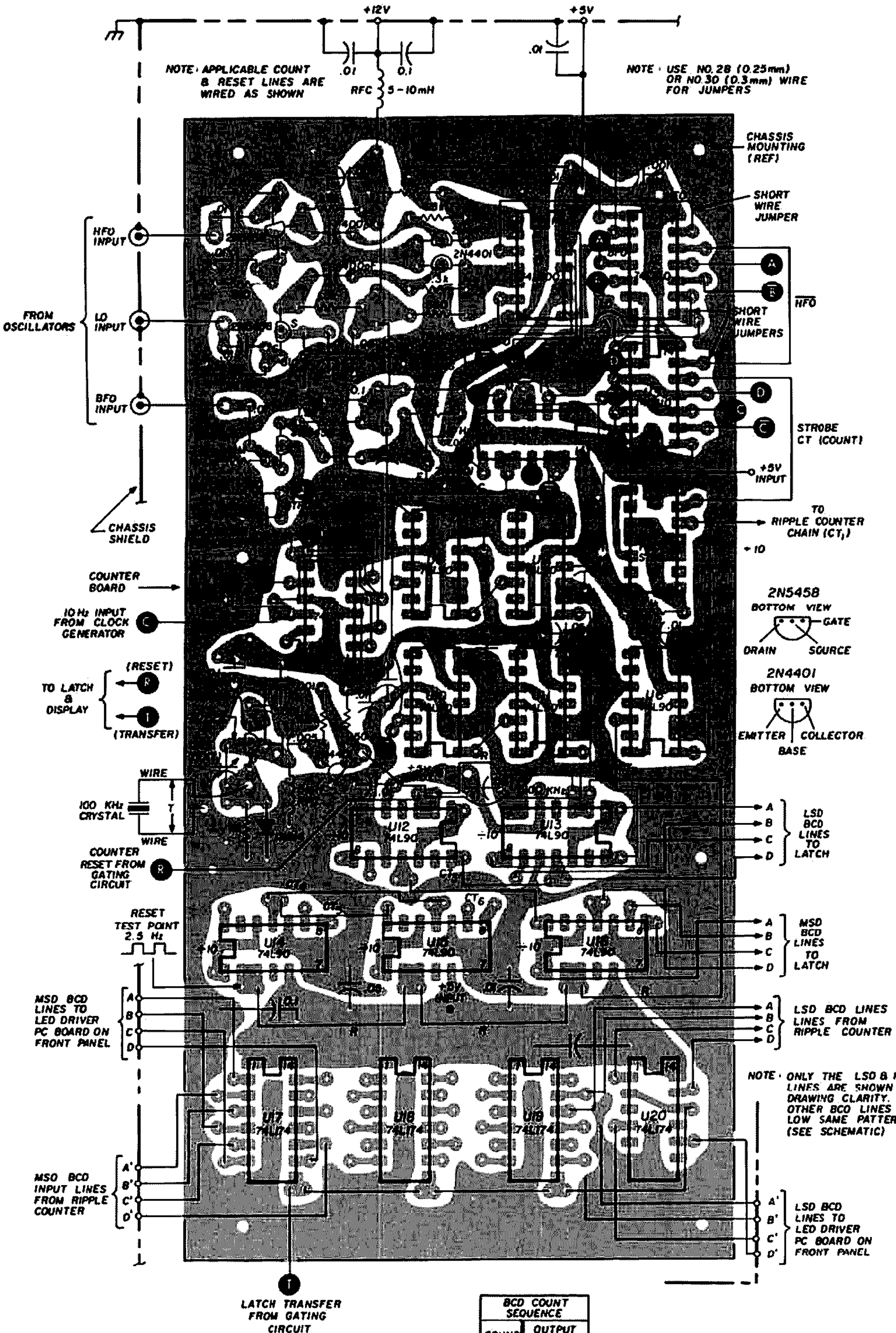
lines and transfer to the output Q lines on the leading edge of the clock, or T pulse in this case. A complete illustration of the latch is shown in **fig. 11.**

In summary, the 14.125-MHz signal is displayed as 14,125 counts in that order. The LED has a decimal point feature so that we can display our received signal as 14.125, directly indicating the MHz and kHz distinctions.

## construction

Construction of the digital counter and display is on two PC boards. One board (**fig. 12**) contains the input buffering, multiplexing, counters, and latch circuits. The second board (**fig. 13**) carries the LED drivers and LED display for convenient mounting behind the front panel. It would have made the construction much simpler to have used a double- or multiple-laminated board; however, when etching PC boards in the kitchen sink, the task of precision art work and registration becomes extremely difficult. The tradeoff here is to use short jumper wires on the component side of the board or use wire-wrap sockets and wire-wrap interconnects.

The LED display is 0.3 inch (7.5mm) high. Easy visibility is obtained from as far as 20 feet (6m). I used a thin sheet of clear plastic in front of the LED display mounted to a bezel (smoked plastic may be used to




NOTE: INPUT BCD INFORMATION FROM THE RIPPLE COUNTER IS TRANSFERED TO THE OUTPUT ON THE POSITIVE EDGE OF THE TRANSFER CLOCK PULSE.

**BCD COUNT SEQUENCE**

| COUNT | OUTPUT D | C | B | A |
|---|---|---|---|---|
| 0 | 0 | 0 | 0 | 0 |
| 1 | 0 | 0 | 0 | 1 |
| 2 | 0 | 0 | 1 | 0 |
| 3 | 0 | 0 | 1 | 1 |
| 4 | 0 | 1 | 0 | 0 |
| 5 | 0 | 1 | 0 | 1 |
| 6 | 0 | 1 | 1 | 0 |
| 7 | 0 | 1 | 1 | 1 |
| 8 | 1 | 0 | 0 | 0 |
| 9 | 1 | 0 | 0 | 1 |
| POSITIVE LOGIC | | | | |

fig. 12. Component placement diagram for the main counter board.

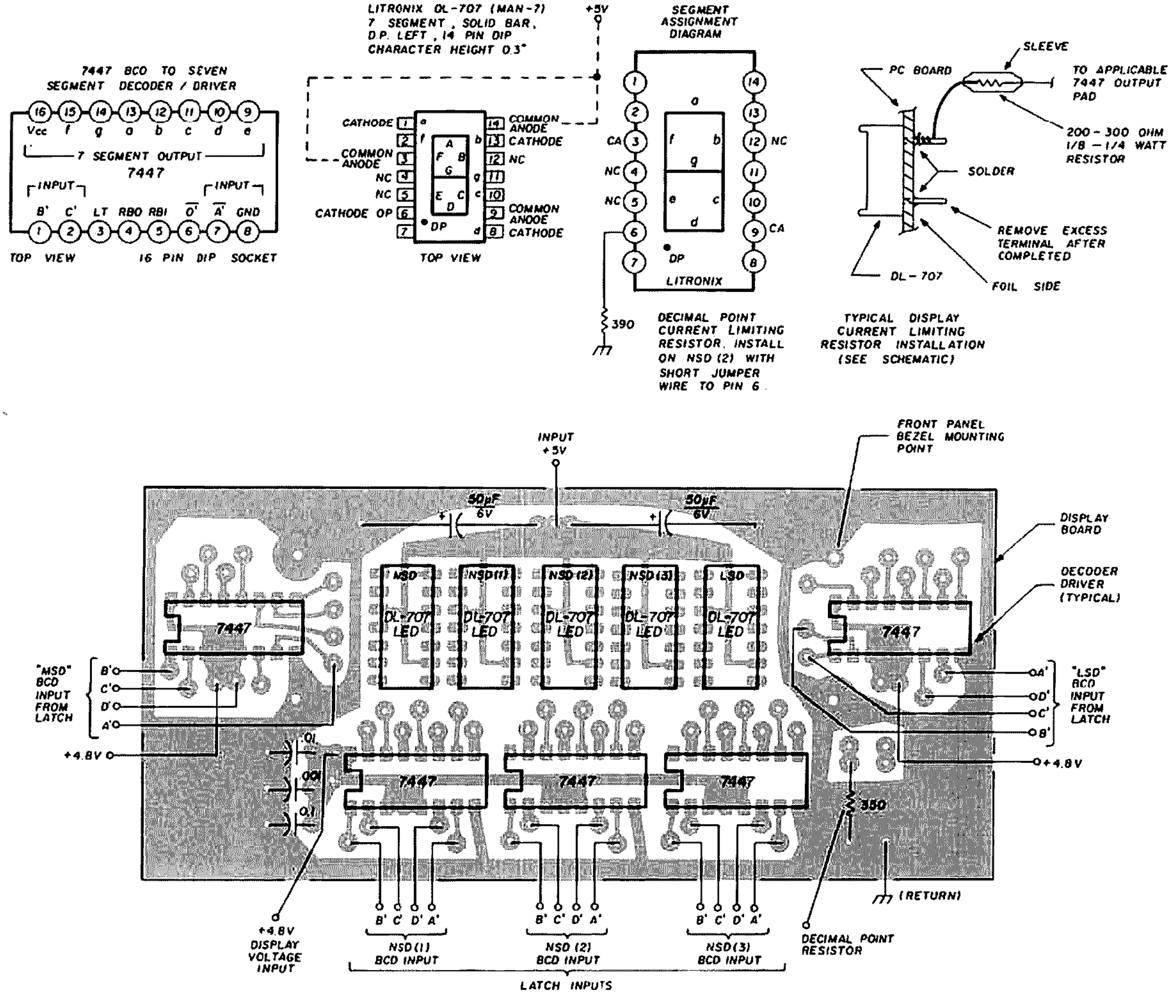



**fig. 13. Decoder/driver and LED display PC component installation.**

enhance the red color wavelength of the diode segments and sharpen up the features).

There are a number of techniques for accurately calibrating the counter. The most obvious is to zero beat the heterodyned 100-kHz clock oscillator against WWV at 5 MHz or against a similar standard reference frequency. Recognizing that the availability of a WWV receiver is limited, I submit the following method as more practical.

My calibration technique was to use the receiver itself to tune WWV to 15 MHz. The HFO and rf tuned circuits are quite adequate for receiving 15 MHz signals.

1. Disconnect the HFO $V_{FV}$ control voltage and, by using an external variable dc supply, the HFO oscillator will tune WWV at approximately 10 Vdc. Peak the RF TUNE and GAIN controls, and use the CW mode position to zero beat the WWV 15-MHz timing pulse, which occurs at 15.001 MHz.

2. Adjust the 100-kHz oscillator resonating capacitor for a 15.001-MHz display.

For a complete description of WWV timing pulse characteristics, you may research your local library or consult a recent edition of the *ARRL Handbook*.

Amateurs in the northwest may find that WWVH or the Canadian Ottawa 14.670-MHz frequency standards are easier to receive. There are several additional European and South American 15-MHz frequency standards that may be used as well. However, I had difficulty in obtaining precise data for the exact position of their timing pulse relative to the carrier resting frequency. WWVH at 15 MHz has a timing pulse whose peak amplitude occurs at 1200 Hz above the resting frequency, which must be considered when adjusting the 100-kHz oscillator for display readout.

The use of an accurately calibrated digital counter to set the 100-kHz oscillator or clock will not be ac-

curate for most applications. An error of ±2 kHz is probably the closest that can be achieved using a secondary counter as a reference. A small amount of LSD toggling will occur, especially when the summation of counts is between 1-kHz intervals. This error in the last digit display is to be expected and is not really bothersome.

## operation

Heat is the biggest single driver for receiver operational stability. Ideally, all the heat-generating devices in the power supply, and heat-sensitive devices in the oscillator circuitry, should be isolated. In practice, however, obtaining an idealized thermal isolation system would be difficult and costly. Using the design and construction approach as illustrated, the normal time constant for good reception stability was on the order of 40 minutes. The entire receiver will take several hours to reach thermal equilibrium. At this point, the frequency drift will be less than 100 Hz per hour. Because of the low power consumption (less than 20 watts), the receiver should run continuously so that the thermal transient effects are minimal.

The thermal time constant will, of course, be a function of the mass and heat-transfer paths as well as characteristic of the chassis, panel, and cabinet construction. A large mass ratio will increase the time constant and minimize the effects of transient thermal changes to the oscillator stability.

Care in power supply heatsinking is also necessary to maintain dissipation and junction temperatures within manufacturer's ratings. Before final alignment of the high-frequency oscillator and display circuits, the receiver should be left on *continuously* for several-hundred hours to settle in various components that have accumulated water hydration during their manufacture, shipment, and storage. This is extremely important if you're interested in readout accuracy.

## reference

1. M. A. Chapman, K6SDX, "High-Performance 20-Meter Receiver with Digital Frequency Readout," Part one, *ham radio*, October, 1977, pages 48.

### appendix

### bandspread techniques

The bandspread capabilities described in part one of this article were limited to the ability of the 10-turn potentiometer that controls the varactor diode voltage for the HFO. Assuming a bandwidth of 350 kHz for both CW and ssb reception, the tuning rate will be 35 kHz per revolution. This tuning rate is not optimal for high resolution CW or ssb reception in crowded conditions. **Fig. A1** illustrates a simple voltage-divider technique for providing about twice the original bandspread capability by dividing the 20-meter band into two parts with a small overlap of about 3.5 kHz in the center. The $V_{FV}$ control potentiometer value is increased by a factor of 10, and

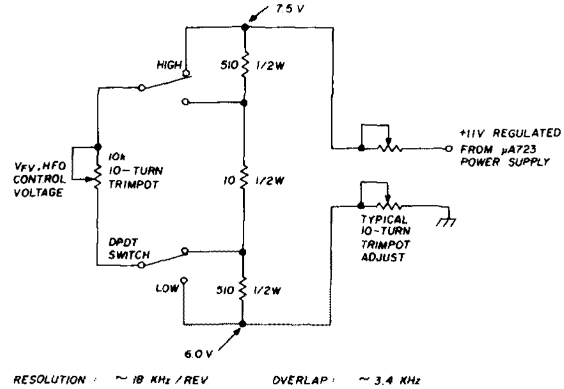



**fig. A1. Split bandspread for high-resolution tuning.**

a series voltage divider network of three resistors plus a simple dpst switch do the rest. Presuming that the original $V_{FV}$ voltage is adjusted for a range of 360 kHz, so that both ends of the band overlap, the new tuning rate is now 18 kHz per revolution of the tuning dial — a significant improvement.

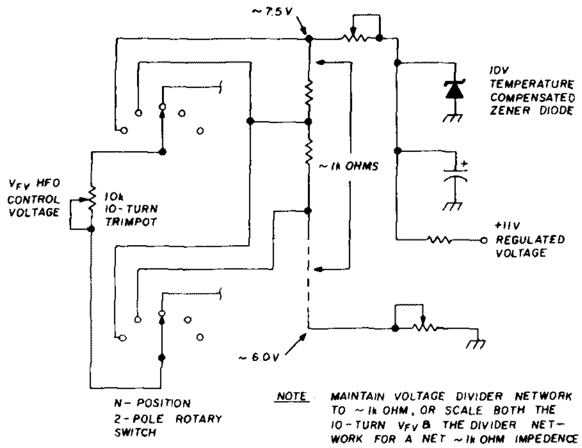



**fig. A2. Ultra high-resolution bandspread scheme.**

**Fig. A2** illustrates an even bolder approach. Using a simple double-pole rotary switch, the band can be divided into any number of segments. A fairly standard switch type is a Centralab 2P rotary switch with 2-6 positions.

Using this approach, the bands could be further subdivided into parts, equal or unequal. Assuming five equal parts at the previous 360-kHz total bandwidth, the potential tuning rate could be reduced to 6 kHz per dial revolution. It should be apparent that, by using various voltage divider arrangements, any desirable bandspread configuration is possible. You could arrange the resistor network for dividing the band into its various operating class ranges or by simply putting emphasis on intervals of most interest.

It is *extremely important* to minimize the thermal heat load on this circuitry due to the high temperature dependency of the components and their resistance value. These parts should be kept away, or isolated from, any high-heat-generating components. Again, to eliminate the change in resistance value resulting from hydration and aging, a burn-in period is desirable before final receiver calibration.

**ham radio**

on the order of 55 pF, while the buffer-amplifier capacitance was only about 25 pF. If the same amount of inductance is used in both stages, a pad of about 30 pF should be used on the buffer (a fixed 22 pF capacitor at U4 pin 1, and about 3-15 pF trimmer capacitance on each inductor). Use 0.02-$\mu$F bypass capacitors on U3, U4 for frequencies below 5 MHz; if all harmonics are to be above 5 MHz, 0.01 $\mu$F will be adequate.

**Meter requirements.** The meter is used to monitor control voltage to the varactor. A meter movement with 100 $\mu$A-1 mA full scale would be a good choice. The multiplier resistor should be adjusted to read 5 volts at midscale. When the op amp is acquiring capture of the vco output, the meter needle will flop from side-to-side as the op amp locks in. Once locked, the meter should again read at midscale; if not, a loss of lock is indicated.

When the op-amp gain-control switch S1, is closed, potentiometer R1 and the 47k resistor, R2, shunt the input-output of U5 (pins 2-6), reducing U5's gain. When gain control R1 is in TUNE, only 47 kilohms remain in the circuit. Thus, the gain is reduced to a minimum, so that optimum capture of the vco signal occurs. When R1 is rotated to the opposite side (S1 still closed), U5's gain will be increased to lock the vco signal. When S1 is opened, maximum gain is available to provide tightest lock control.

Bias-set potentiometer R3 should be a small, multiturn trimpot. This control is extremely critical. It must provide an indicated voltage on U5 pin 6 that is *exactly* the same as the 5-volt supply voltage reading. Less than one-quarter turn of this control will cause the meter needle to flop from one side to the other as the bias is being set. Only at the proper setting will U5 have maximum control of the varactor and thus lock the vco. Once set, R3 should require no further adjustment unless a circuit revision is made; therefore, this control should be mounted on the rear of the circuit board out of reach of accidental change.

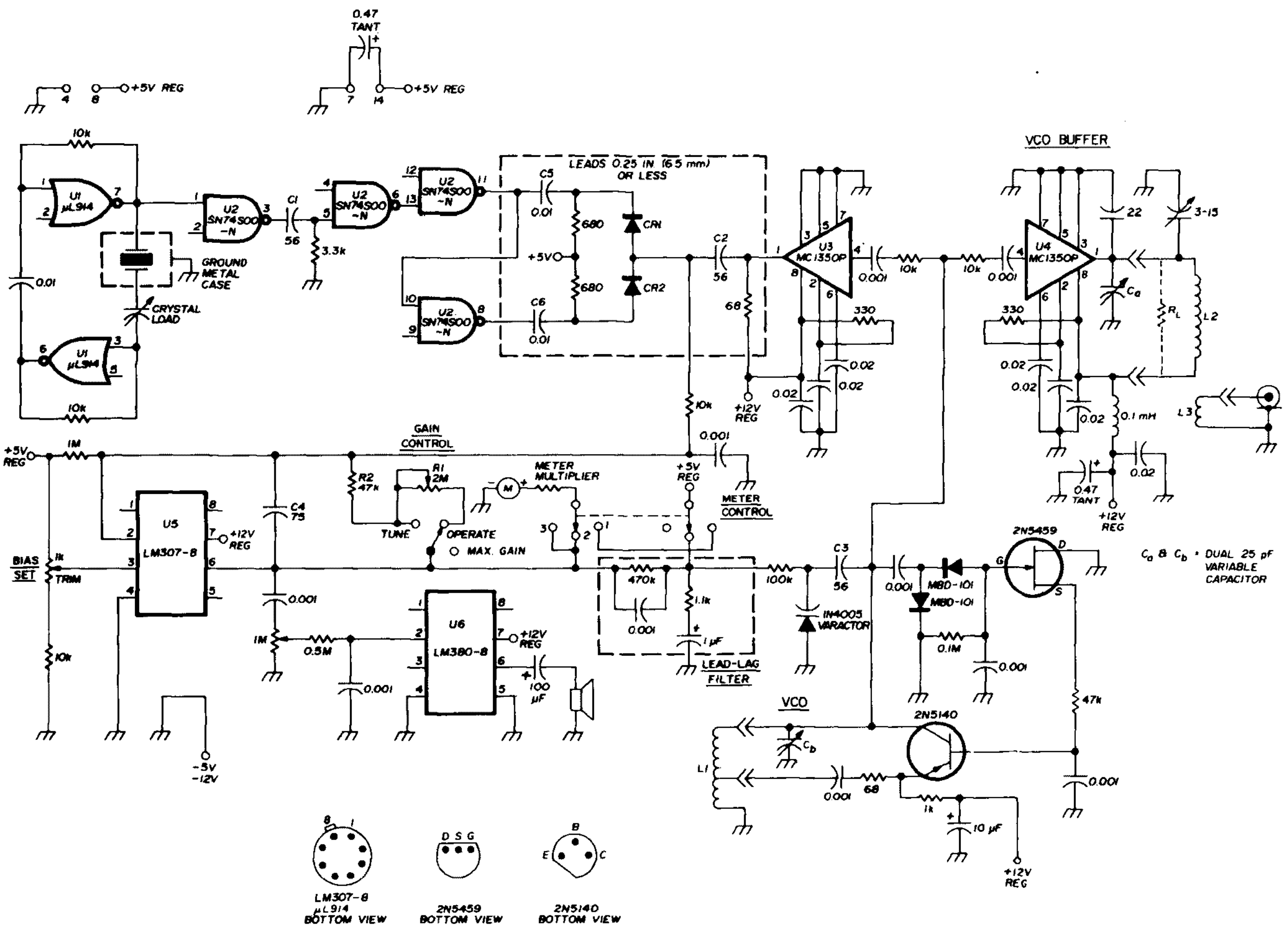


fig. 1. Schematic of the phase-locked loop, crystal-controlled harmonic generator. Circuit provides high-accuracy integer harmonics from a basic crystal-oscillator frequency.

**Capacitors.** Several of the capacitors in the circuit should be of good quality. Capacitor C1 in the Schottky network; capacitor C2; and capacitor C3 should be dipped mica components. Capacitor C4 should be a silver mica. Capacitors C5, C6 should be small 50-volt 0.01-$\mu$F disc ceramics with short leads.

With the exception of tantalum capacitors marked on the schematic, and electrolytic capacitors (marked +), all other capacitors may be disc ceramics. (If identical rf output is required on all harmonics, the vfo-buffer output inductors may require loading resistors; 1-2k should suffice.)

**Power supply.** The regulated 5- and 12-volt supplies can be built with transformers that have outputs of 8 and 16 volts, respectively. **Fig. 2** shows a regulated supply using one center-tapped transformer secondary.

**Diodes.** All silicon diodes show some variable capacitance with reverse bias. The varactor I used is a 1N4005 and has better $Q$ than some diodes designed for varactor use. Referring to **fig. 3**, measure the vco frequency at 2-, 5-, and 8-volts reverse bias. If the frequency change from that of the 5-volt reading is greater than 0.8% of center frequency, the diode will be satisfactory as a varactor. Several diodes I tested showed good yields: the 1N4005 was about 50%; the 1N4007 about 20%, and the Motorola HEP 170 (four tested) showed over 1% frequency change at 5-MHz center frequency.

## tuning and adjustment

A frequency counter that covers the desired frequency range is a decided asset. Aligning the vco and buffer frequencies is much more easily accomplished with a counter. If a counter isn't available, a continuous-coverage communications receiver can be pressed into service.

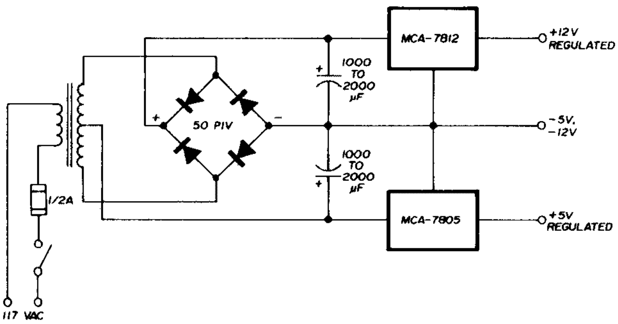



**fig. 2. Suggested power-supply circuit, which provides regulated 5 and 12 volts.**

Assuming the crystal oscillator is working and that the vco is oscillating near one of the required harmonics, use the following steps.

**1.** Set the meter switch (**fig. 1**) to position **1** and set gain-control pot R1 to maximum resistance (switch closed).

**2.** Note the 5-volt power-supply voltage reading.

**3.** Set meter switch to position **2** and adjust the bias-set pot on the op amp until the meter reads *exactly* the same as the reading in step **2**. As noted previously, this bias-set pot control is extremely touchy.

**4.** Adjust the trimpot control until the meter again reads the same as the 5-volt supply reading.

## operation

**1.** Shift the meter switch to position **3** for operation. The meter will now show the control voltage impressed onto the varactor under normal operating conditions.

**2.** Set the gain-control pot to minimum resistance (minimum gain) and tune in a harmonic to zero beat. Integer harmonics will be quite loud, while others will be weak.

**3.** Determine the harmonic frequency and adjust the components until the required harmonic output is obtained.

**4.** Increase gain and note the tighter lock that occurs. Increasing gain until the switch opens provides maximum gain. As a trial, note how sharply the audio signal of the heterodyne goes through zero beat when gain is minimum; also note the dial divisions that take it above audibility.

**5.** Now increase the gain to achieve lock and note how far the vco dial can be moved before lock is lost. Also note that the meter will go to one side or the other of midscale, which shows that the op amp is trying to hold the vco in lock by changing the varactor's reverse bias. If the meter needle suddenly flops to one side, you have lost lock.

**6.** Adjust the each inductor in the vfo and buffer to tune all harmonics with the capacitance previously mentioned. When all harmonics have been calibrated, log the switch and vco dial settings for each.

## conclusion

The crystal-controlled PLL vco will provide high-accuracy integer harmonics from a basic crystal frequency. If frequencies to 50 MHz or more are required, I suggest a crystal frequency of at least 1 MHz unless closer channel spacing is required. If harmonics spaced 2, 3, or more MHz are required, consideration should be given to using a crystal with a fundamental frequency equal to the channel spacing

required. However, if a lower channel spacing or lower frequency is the requirement, use a lower-frequency crystal or use a divide-by-10 and/or divide-by-n combination to arrive at the desired channel spacing.

If the final output frequency is very high and the channel spacing is very low, such as in two-meter

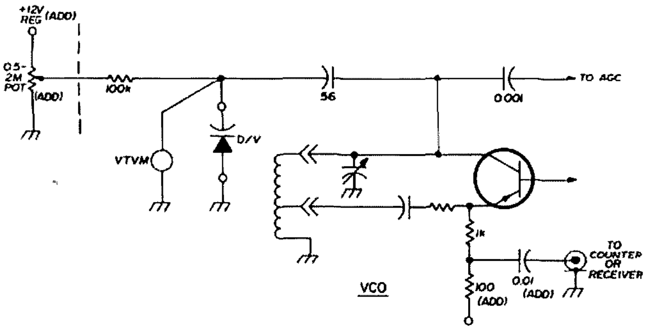



**fig. 3. Test setup *using* the vco to determine suitability of diodes for varactor service.**

work, generate the required channels as stated, then mix upward with a high-frequency crystal to arrive at the required output frequencies.

If the requirement is high accuracy, I recommend the use of a *high-accuracy* crystal. These are available with guaranteed accuracies on the order of a few parts per million. If used in a temperature-controlled oven, higher accuracies may be obtained.

The unit built by W4OQ uses a 1.000 000-MHz crystal* which supplies harmonics from 6-36 MHz for use as a local oscillator. This local oscillator will be mixed with a Collins PTO to provide a solid-state, continuous-coverage signal generator from 1.5-33 MHz. When compared with WWV at 15 MHz, the PLL vco showed −10 Hz at turn-on, −5 Hz after 5 minutes, and −3 Hz after 10 minutes. It went through zero error at 35 minutes. After two hours, it showed variations not in excess of 2 Hz — not bad for a unit without an oven or elaborate temperature compensation!

The unit draws only 30 mA from the 12-volt supply and 25 mA from the 5-volt supply. The rf output is approximately 150 mV rms into a 50-ohm load.

*International Crystal Co. Type H-A.

## references


1. K. W. Robbins, W1KNI, "Transistors and ICs in a Phase-Locked Local Oscillator," *QST*, January, 1972. (Corrections in QST, February, 1972 and subsequent issues of *The ARRL Handbook*.)
2. K. W. Robbins, W1KNI, "PLL Update," July, 1975 (unpublished).


**ham radio**

# improved receiver selectivity and gain control

## An alternative to the manually switched front-end attenuator for hf receivers — a PIN diode attenuator and cascaded i-f filters

**Once upon a time** there was a receiver called the AR-88, which used cascaded rf amplifiers and a mixer that was far from crunchproof. People would listen to the signals from the receiver (many of which were not actually in the band) and wonder at the sensitivity of this receiver.

And it came to pass that high-frequency i-f filters were invented, and double-conversion went the way of the dodo — receiver designs were changing. Solid-state front ends still generated distortion products, so the antenna attenuator was invented, thus creating a plethora of twitchy index fingers. Wise men said, "Ye shall not agc the front end, for operating point changes causeth dynamic range to suffer." And so the attenuator remained with us.

Then came Doug DeMaw,[1] and certain German gentlemen and wise men from California. They produced crunchproof mixers saying, "Lo, the rf amplifier is no more, and we suggest cascading i-f filters besides." The world marvelled, hand on attenuator switch, awaiting developments. It was with such history in mind that the following ideas were born, while updating some previous receiver designs.[2,3]

The antenna attenuator is a great idea, just in case the crunchproof mixer is less than ideal — but that front panel switch just had to go. A PIN diode rf attenuator, agc controlled, was the alternative providing many dB of attenuation ahead of the mixer across the entire range of an hf receiver.

Previous experience[2] showed that cascading i-f filters, instead of merely switching them, resulted in greatly improved adjacent-channel selectivity — where it really counts. To change bandwidths, shorting the sharper filter seemed to be simple and effective.

Finally, all my previous receiver efforts[2,3,4,5] seemed to suffer from inadequate agc control range. Nothing short of 60 dB or better seemed satisfactory, so this range was set as a target number.

The following circuitry is the end result of these deliberations — another "great leap forward" in the endless quest for the ultimate in homebuilt receivers.

### circuit description

The PIN diode attenuator (**fig. 1**) is designed to be inserted between the antenna and antenna input connector of any hf receiver. The PIN diode has a very low impedance when conducting a relatively high bias current and a very high impedance when the bias current is small. While most PIN diodes are designed to be used above 100 MHz, certain Hewlett Packard diodes are useful down to 1 MHz. The attenuator is built in a separate shielded enclosure, with coaxial connectors provided on each end. Depending on construction, this attenuator should provide up to 40 dB attenuation (and, therefore, agc range) between 3-30 MHz when terminated in a 50-ohm impedance.

The PIN current source, buffer, and agc circuits are built on a separate board or chassis. An npn transistor is used as a current source, providing more than 100 mA to the PIN diode. The current-source transistor is driven from the agc circuit through a jfet buffer, Q3, which prevents the low impedance of

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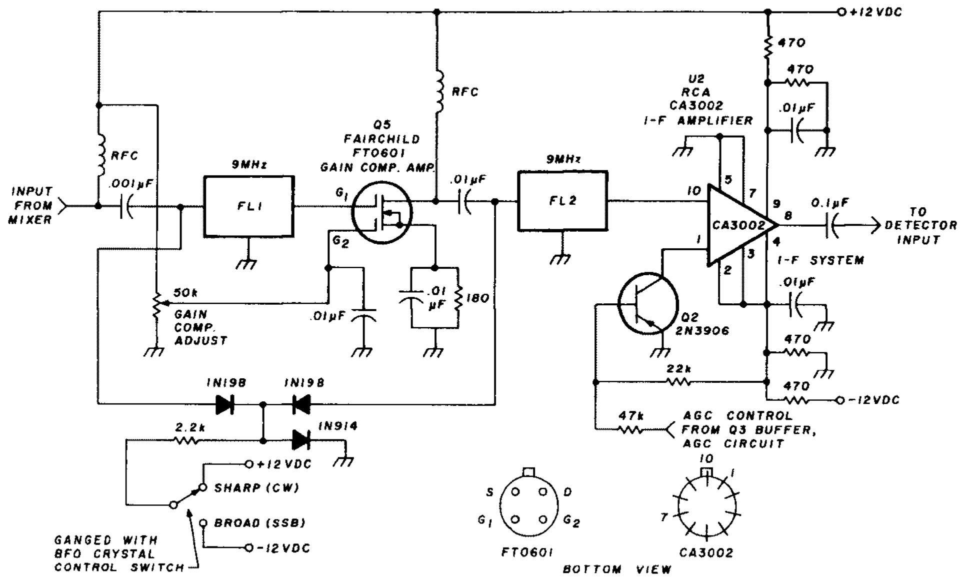
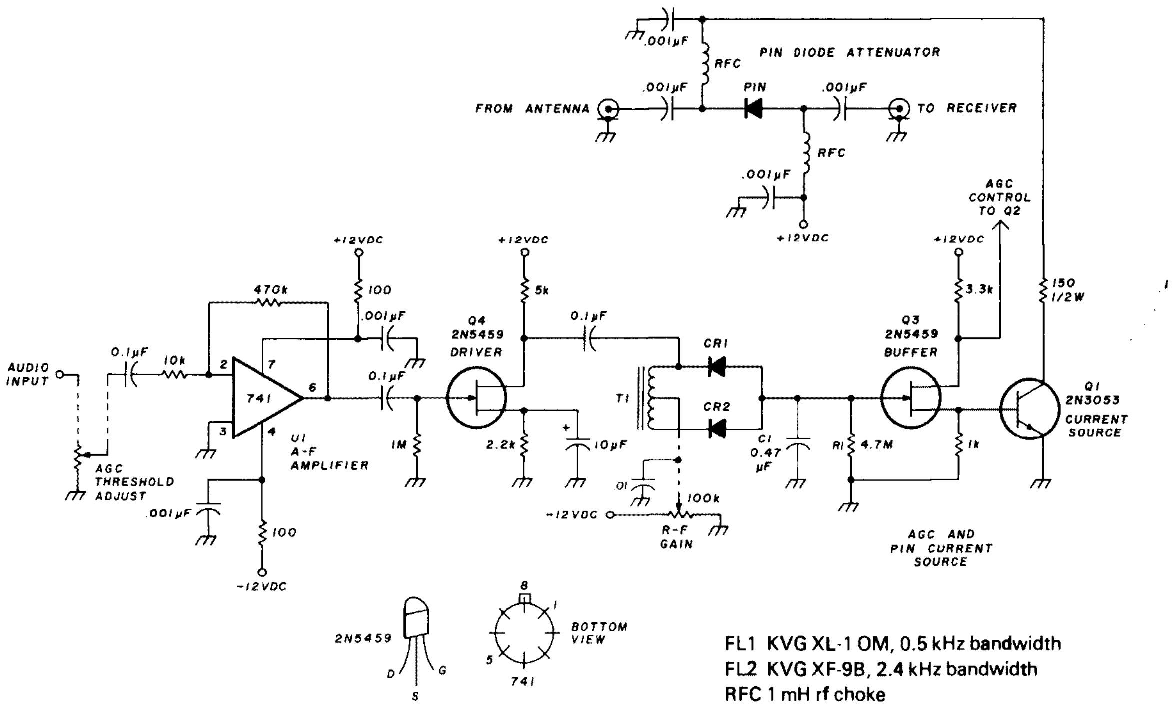



| | | | |
|---|---|---|---|
| CR1, CR2 | Germanium diode | RFC | 1 mH, 100 mA rf choke |
| PIN | Hewlett-Packard 5082-3379 | T1 | Any small audio transformer |

**fig. 1. The PIN diode attenuator, A, and Pin-diode current source, buffer, and agc circuits, B. The attenuator should provide up to 40 dB attenuation between 3-30 MHz when terminated in a 50-ohm impedance.**

current source Q1 from loading the agc line and affecting its time constant. This time constant is determined by R1 and C1; values shown are a compromise between slow (ssb) and fast (CW) agc.

The agc voltage is audio-derived; audio from the top of the receiver audio gain control is amplified and rectified, with 200 mV rms at the input of U1 sufficient to cause maximum attenuation. The centertap




FL1 KVG XL-1 OM, 0.5 kHz bandwidth
FL2 KVG XF-9B, 2.4 kHz bandwidth
RFC 1 mH rf choke

**fig. 2. I-f system uses cascaded filters. When used with the PIN-diode attenuator an agc range of more than 70 dB was measured. FL1 is a KVG XL-1 OM, 0.5 kHz bandwidth filter. FL2 is a KVG XF-9B 2.4 kHz filter. The rf choke is 1 mH.**

of T1 can be grounded. However if it's tied to the wiper of a potentiometer connected between ground and −12 Vdc, manual control of attenuation level, while maintaining the automatic feature, provides an rf gain control function for the receiver.

The i-f strip (**fig. 2**) uses an RCA CA3002, which provides 30 dB gain at 9.0 MHz and is specified as having 80 dB of agc control range. Used in conjunction with the PIN attenuator, agc range of more than 70 dB could be measured. If the spec-sheet people are honest, it should be around 120 dB, far beyond most instrumentation measurement capability.

The agc control pin of U2 is driven through Q2 from the PIN attenuator agc system. Circuit values are used that ensure only nominal attenuation by U2 until the attenuation limits of the PIN attenuator are approached; after this point, attenuation in the i-f IC increases very rapidly.

In front of the i-f chip, U2, two KVG 9-MHz crystal filters are installed so that in normal operation the two filters are cascaded. The sharp filter (500-Hz bandwidth) drives a mosfet amplifier, Q5, which has its own gain control. This amplifier is adjusted to have a gain that exactly compensates for the insertion loss of the sharp filter. This ensures that the i-f strip will have the same gain whether or not the sharp filter is in circuit. When a wider bandwidth is desired, the sharp filter and its compensation amplifier are simply shorted by activating a diode gate. The switch that controls the diode gate should also switch in the proper bfo crystal when the sharp filter is in use. The input circuit of the i-f board is designed to supply the dc operating voltage to the mixer which drives it.

Only the circuit layout of the PIN diode attenuator is critical. Attenuation can be compromised by stray capacitance, so all leads should be as short as possible. Only disc ceramic capacitors should be used. The attenuator uses a section of PC board as a chassis with ground connections soldered right on to the copper foil. The attenuator enclosure was also made from copperclad board. The entire assembly was soldered together after the final tests were completed.

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2. Mike Goldstein, VE3GFN, "A Solid-State 80-meter Receiver," Radio Handbook, 19th edition, Editors and Engineers, page 20.15.
3. Mike Goldstein, VE3GFN, "An Advanced Solid-State Deluxe Amateur-band Receiver," Radio Handbook, 19th edition, Editors and Engineers, page 20.23.
4. Mike Goldstein, VE3GFN "A Portable Amateur Band Receiver," Radio Handbook, 19th edition, Editors and Engineers, page 19.20.
5. Mike Goldstein, VE1ADH and George Cousins, VE1TG, "High Quality Hybrid Receiver," *73*, February, 1968, page 42.


**ham radio**

# receiver spurious response

## and its cures

Spurs are
for cowboys —
not amateur
communications receivers!

**Recent literature has paid a good** deal of attention to intermodulation distortion in receivers; this is one form of spurious signal generation but many others are possible. These sources can be improper frequency sets in mixers, overdriven amplifiers, digital circuits, parasitic oscillations, and inadvertent coupling. Whether in new designs or modifications, all can be eliminated or at least minimized.

Mixers are the most probable spur sources. Since a mixer must be nonlinear it will generate harmonics internally. A poorly chosen set of input frequencies can produce spurs or birdies at the output. Intermodulation distortion spurs have been well covered in the past, so attention is directed to identifying good and bad frequency sets.

A lot of methods and charts, including an HP-25 program, have been generated.[1-3] One of the very simplest and easiest to use is that of Fisk[4] and involves finding the decimal ratio of the lower input frequency set divided by the higher input frequency set; a similar method was described by Stevens.[5]

### decimal ratio method

Division of lowest frequency set by the highest will always result in a number between zero and one. This quotient can then be used to identify good and bad sets with appropriate tables. It does not matter which input has what; either may have the lower frequency ($F_L$) or the higher frequency ($F_H$); either or both inputs may have a single frequency.

**Tables 1** and **2** show the ratio, spur-product at that ratio, spur order. The rightmost columns relate either $F_L$ or $F_H$ to the mixer output frequency at each ratio. **Table 1** is for difference outputs ($F_H - F_L$) while **table 2** is for sum outputs ($F_H + F_L$).

The sum of each spur product frequency multipliers is the order of the spur amplitude. A lower order produces the stronger spur. Maximum spur order has been limited to 6. Higher orders are a problem only in precision instrumentation.

**Example**. The conventional broadcast receiver tunes 550 to 1650 kHz with a 445 kHz i-f; the local oscillator range must be 1005 to 2105 kHz and reasonable input selectivity is assumed.

Spur analysis will use **table 1** since mixer output is the difference. Decimal ratios will be $F_L/F_H = 550/1005 = 0.547$ at the low end and $F_L/F_H = 1650/2105 = 0.784$ at the high end. Three spurs occur:

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| ratio | spur product | order | difference, terms of lower frequency | zero beat |
|---|---|---|---|---|
| 0.600 | $4F_L - 2F_H$ | 6 | $2F_L/3$ | 682.5 kHz |
| 0.667 | $2F_L - F_H$ | 3 | $F_L/2$ | 910.0 kHz |
| 0.750 | $3F_L - 2F_H$ | 5 | $F_L/3$ | 1365.0 kHz |

The *zero beat* column is found by equating the i-f to the *difference, terms of lower frequency* column. This allows identifying where the birdie is located on the dial.

Only the 910 kHz birdie will be found when the signal is strong, provided the rf stage (if used) is linear. Why the strong signal requirement? The nonlinear amplitude characteristic of the mixer produces less harmonic output at lower signal levels so there is a difference as to which frequency set has the stronger signal. An LO set of 95 to 1195 kHz (if possible) would still produce a 0.667 ratio but the spur would be more pronounced since $F_H$ (1365 kHz) is the fundamental and the antenna input signal.

## predicting spur levels

The variety of circuits and different nonlinear characteristics makes it difficult to predict spur

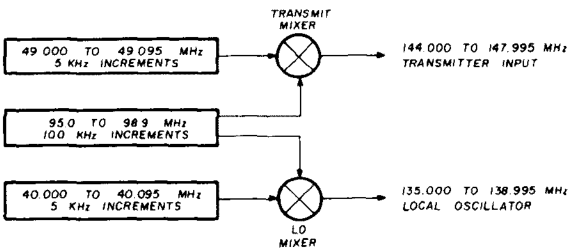


**fig. 1. Mixing synthesizer frequency control system for a two-meter transceiver with a 9-MHz i-f. This system has several unwanted spurs, as discussed in the text.**

levels. Input power level differences vs $F_L$ and $F_H$ also make it difficult. The intercept point concept explained in a recent *ham radio* article[5] is a great help for estimating spur level if the intercept point data is available.

Some integrated-circuit mixers such as the Texas Instruments SN76514 include typical spur product levels. If in doubt, a breadboard test with a frequency set known to produce a low-order spur is helpful. A spectrum analyzer is not required since a receiver with a calibrated attenuator will serve as well. Frequencies are chosen so that outputs are slightly separated between desired and spur-produced signals.

Lacking either data, spur-product multipliers should be examined and one input should be low. Spurs are down with greater level differences.

Balanced mixers will have a reduction of even harmonics. Double-balanced types using transformer coupling are designed principally for input-output isolation; they reduce even harmonics from the source, not those created internally.

External harmonics will increase the spur product level. If spur products are unavoidable, the higher level input should be as clean as possible.

The best way to avoid mixer spurs is to pick the right frequency set. Keep the order or harmonic number as high as possible if some spur ratios exist; select frequencies while still in the block diagram stage.

## starting a transceiver design

A two-meter, incremental tuning transceiver is to be built with a 9-MHz receiver first i-f. A mixing synthesizer is first considered for frequency control; the block diagram is shown in **fig. 1**. The method of generating each frequency set may be put aside until a spur check is made.

Each mixer upconverts to each output so **table 2** is used for the check. Each frequency set is variable so the ratios have two limits: lowest frequency of one input vs highest frequency of the other, then the highest frequency of the first input vs the lowest of the other.

The LO output ratios are then 40.000/98.9 = 0.404 and 40.095/95.0 = 0.422. No problem. Transmit output ratios are 49.000/98.9 = 0.495 to 49.095/95.0 = 0.517 which crosses the 3rd order spur listed in **table 2**. Identification of zero beat is done by checking the 3rd harmonic of the $F_L$ set for equality with 1.5 times the $F_H$ set. Here it occurs at 147.000 MHz with birdies adjacent.

A possible spur source is direct coupling of the transmit side lower frequency set 3rd harmonic bypassing the mixer. These range from 147.000 to 147.285 MHz, all in-band. Shielding and supply decoupling have to be considered.

By altering the frequency set ranges, spurs and direct harmonics can be moved out of band. The alteration would be:

**table 1. Spur identification for difference output.**

| $F_L/F_H$ ratio | spur product | spur order | difference output due to spur product in terms of: $F_L$ | $F_H$ |
|---|---|---|---|---|
| 0.143 | $6F_L$ | 6 | $6F_L$ | $6F_H/7$ |
| 0.167 | $5F_L$ | 5 | $5F_L$ | $5F_H/6$ |
| 0.200 | $4F_L$ | 4 | $4F_L$ | $4F_H/5$ |
| 0.250 | $3F_L$ | 3 | $3F_L$ | $3F_H/4$ |
| 0.333 | $2F_L$ | 2 | $2F_L$ | $2F_H/3$ |
| 0.400 | $4F_L - F_H$ | 5 | $3F_L/2$ | $3F_H/5$ |
| 0.500 | $F_L$ | 1 | $F_L$ | $F_H/2$ |
| 0.600 | $4F_L - 2F_H$ | 6 | $2F_L/3$ | $2F_H/5$ |
| 0.667 | $2F_L - F_H$ | 3 | $F_L/2$ | $F_H/3$ |
| 0.750 | $3F_L - 2F_H$ | 5 | $F_L/3$ | $F_H/4$ |

| | |
|---|---|
| Lower transmit set | 32.000 to 32.095 MHz |
| Common higher set | 112.0 to 115.9 MHz |
| Lower LO set | 22.000 to 22.095 MHz |

Spur checks now give ratios of 0.264 to 0.276 on the transmit side and 0.190 to 0.197 on the LO side. No ratios fit any spurs, and harmonics of direct coupling are not in-band. The latter includes cross-coupling from transmit side to LO side or vice versa.

Even though no spurs exist, this may not be attractive from a system standpoint. We can try a single variable set upconverted with fixed frequencies.

This is shown in fig. 2 with the single variable set derived by a phase-locked-loop or manually-tuned oscillator. The single crystal oscillator at 4.5 MHz can be counted down for the PLL reference; it must be shielded to prevent second harmonic interference with the i-f.

A spur check shows the transmit side is okay with ratios of 0.600 to 0.644. The LO mixer ratios are 0.667 to 0.716 with a 5th order spur at 135 MHz (144.000 MHz input). This results from $3F_H - 2F_L$ or $4F_L - F_H$. It may be at the band edge but a birdie is possible at 144.005 MHz.

The next consideration is the common lower frequency set. Ratios will be 0.200 to 0.289 with a 6th order spur at 54 MHz due to $6F_L$ or $2F_H - 4F_L$. This is again at 144 MHz. If a PLL is used, it should have lowpass filtering or a square-wave output to minimize even harmonics.

It is difficult to avoid spurs in a single-crystal scheme if a harmonic of the crystal falls in-band. Fortunately, the spur orders are high with low spur product levels and both land on the band edge frequency. (Note that the band edges should be locked out with PLL control logic.)

The circuit of fig. 2 is favored for simpler structure and adapability to phase-locked-loop control of a single variable frequency set. It does have some spurs and possibility of i-f interference so the oscillator and multiplier power levels should be low. If power is needed, amplifiers and filters should be added after the mixers. In addition, the PLL is basically a digital device so some thought should be given to harmonics from such sources.

**table 2. Spur identification of sum output. Note that there are two possible spur products for each $F_L/F_H$ ratio.**

| $F_L/F_H$ ratio | spur products | | spur order | sum output due to spur product in terms of: $F_L$ | $F_H$ |
|---|---|---|---|---|---|
| 0.200 | $6F_L$ | or $2F_H - 4F_L$ | 6 | $6F_L$ | $6F_H/5$ |
| 0.250 | $5F_L$ | or $2F_H - 3F_L$ | 5 | $5F_L$ | $5F_H/4$ |
| 0.333 | $4F_L$ | or $2F_H - 2F_L$ | 4 | $4F_L$ | $4F_H/3$ |
| 0.500 | $3F_L$ | or $2F_H - F_L$ | 3 | $3F_L$ | $3F_H/2$ |
| 0.667 | $4F_L - F_H$ | or $3F_H - 2F_L$ | 5 | $5F_L/2$ | $5F_H/3$ |

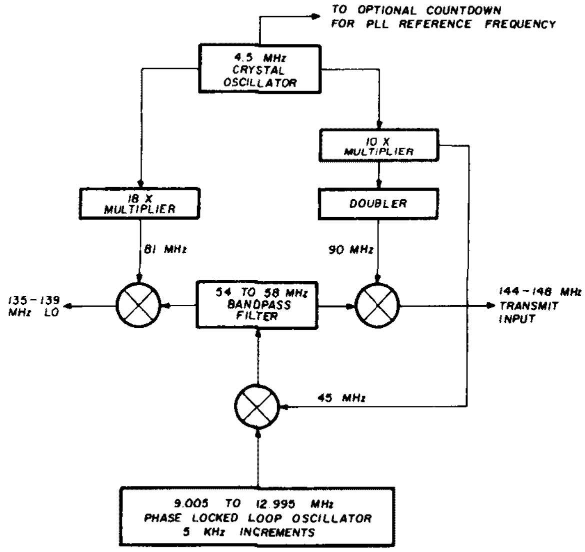


**fig. 2. A 144-MHz transceiver frequency control circuit using a phase-locked-loop oscillator and mixing with a crystal oscillator reference.**

## digital source harmonics

Assuming rise and fall times are equal, harmonic voltage magnitude can be calculated relative to peak waveform amplitude at video with

$$e_n = \left[\frac{2A}{n^2\pi^2 F t_r}\right] \left| Sin\,(n\pi F\, t_r)\; Sin\,(n\pi F\, t_w)\right| \quad \textbf{(1)}$$

Where:

$e_n$ = Peak harmonic voltage at nth harmonic
$A$ = Peak video waveform amplitude
$n$ = Harmonic number
$F$ = Waveform frequency, MHz
$t_r$ = Rise or fall time, microseconds
$t_w$ = Half-amplitude video pulse width, microseconds

Both sine arguments are in radians and absolute sine values are used. This is a simplification of the Fourier coefficient formula for equal rise and fall times.[7]

A video waveform of width equal to half period will have only odd harmonics. Magnitude of the width sine term will be unity with $n$ odd, zero with $n$ even. Such simplification is all right for rough calculations, but don't expect digital ICs to be that perfect. It is better to make worst-case calculations by offsetting a square-wave by half the data sheet rise or fall time. This is only part of the digital source harmonics.

## supply line transition surges

TTL and CMOS digital ICs have a current surge

every time an output changes logic states. This surge is only nanoseconds wide so the harmonic content reaches the uhf region. Current surge ampltude varies between devices and examples may be found in the manufacturer's data books.[8] Many digital designers ignore this RFI source with limited supply bypassing. These glitches seldom affect the digital circuit itself so it's not their worry; it can be murder to receiver sensitivity.

A rule of thumb is to provide a 470 to 1000 pF bypass capacitor for every output pin on the IC, using disc ceramics with the shortest possible leads, installed right at the IC package. A good mounting location is on the foil side directly between $V_{CC}$ and the ground pins. This may seem like overkill in bypassing, but try a receiver with a "sniffer" loop at the digital supply lines of a limited bypass circuit.

Generally, CMOS logic is quieter in surges with low supply voltages. Schottky TTL is quieter than low power TTL even though the speed-power product is better. Line driver or buffer packages have the highest surges.

## digital source interfacing

Loose capacitance coupling is seldom a problem to a digital output. Such capacitance is usually less than the device test load. Rise and fall times of CMOS with capacitance loading will be nearly equal but TTL is different. A standard TTL totem pole output can sink 16 mA at logic 0 (+0.4V), but will only source 0.4 mA at logic 1 (+2.4V minimum). A high to low transition will always be faster with loading.

It's another story to drive a filter. Drivers for 50- to 75-ohm lines are available but an open-collector gate output as in **fig. 3** is less expensive. In this circuit, the collector resistor is the filter source impedance (approximately) and a 470-ohm value fits medium speed and low power Schottky TTL gates. All other outputs for purely digital use must also have pull-up resistors.

## digital mixers

The technique of digital mixing has been largely unexplored. Any 2-input gate can mix two digital sources. Because of fast transitions, harmonic content is high so strict observance of spur products must be made. It is extremely simple and an open-collector 2-input NAND gate could serve as the PLL mixer in **fig. 2**. In this circuit the 45-MHz analog signal would require either a fast comparator or Schmitt-input inverter or gate as an interface. Schottky or high power TLL devices are required at such frequencies. A different frequency set is needed since the spurs at the lower band edge will be more pronounced.

Digital mixers have the advantage that spurs are more easily predicted mathematically although a computer program may be necessary. It is a scheme open for experimentation.

## zero i-f ssb receiver

The direct-conversion receiver has become popular for 40- and 80-meter work; a digital counter is invariably used to provide the necessary

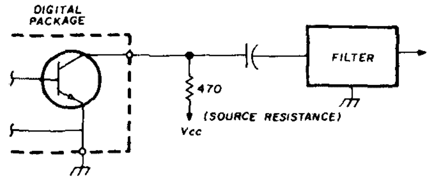


**fig. 3. Digital to analog interface through a filter using open-collector outputs.**

*quadrature* LO phase. Unfortunately, a 40-meter receiver becomes susceptible to 15-meter pickup.

The cause is the counter for LO injection. A 20 nanosecond rise/fall time square-wave will have a third harmonic only 10 dB below the fundamental in voltage. This is high enough to provide mixing action although more conversion loss exists.

Adequate antenna input filtering is required. LO filters can't be used since this destroys the quadrature relation. This is a good example of source-related spurs.

## spur suppression when not in zero beat

Close birdies can be reduced by an active limiter interface such as the circuit from a broadcast fm i-f. Since a limiter is an amplifier driven to cutoff and/or saturation, the stronger the input, the more a low-level input is attenuated at the output; *i.e.*, the low signal can't get through unless the amplifier is in the linear amplitude level.

A relatively low cost linear is shown in **fig. 4**. The metal-can version MC1590 can be substituted in this circuit since it has the same chip as the MC1350. These Motorola devices are wideband, differential input and output amplifiers that also work well as limiters into the 50-MHz region.*

Ac input coupling is required. The input impedance is a constant 5k in parallel with a 5 pF at each input. Each output is an open-collector to allow a variety of loads and maximum voltage swing.

*Diode limiters will work well if the forward voltage of each is matched, but diodes absorb power. The object of use is as a combination buffer and suppressor with one mixer input level low to minimize spurs. The circuit shown has been test flown in three different avionics systems.

Transformer coupling the output reduces even harmonics. The input may be single-ended or push-pull with little output difference. These devices also make good mixers.[9]

## ground loop ghosts

Hours can be spent hunting a ground loop that isn't there. This usually occurs in a wire bundle carrying both high and low level signals, and coupling can occur even when both lines are coax. Why?

Woven braid outer conductor coax has only 98 per cent shielding, at best. This can drop to 90 per cent at sharp corners. The result is little coupling apertures all along the line with greater coupling at higher frequencies. One cure it to use double-braided coax; a cheaper method is to wrap each line with aluminum foil. For even better results, separate the high and low level lines.

## other sneaky paths

A beautifully shielded source may be a marvel of mechanical work but it is all window dressing if the supply lines aren't filtered. Watch out for bypass capacitor resonances; they can become inductors above resonance and lose all effectiveness. Parallel small, medium, and large value capacitors when in doubt.

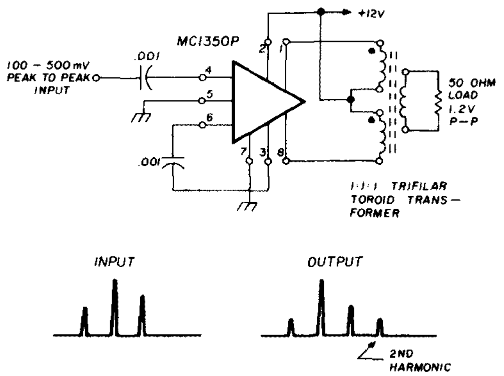


fig. 4. Using an active limiter as a spur suppressor. The input and output signal spectra are shown at the bottom with spurs to either side of the center frequency (logarithmic vertical scale).

Apertures become antennas at higher frequencies. A receiver with a sniffer loop will help find the cause. If the aperture can't be easily closed, try aluminum foil held in place by a thin layer of contact cement.

Window screen material is fairly good for shielding and allows air circulation. One caution: make certain it is metal; a lot of screen material is plastic these days!

Control lines in the dc/audio range can be antennas, so watch the routing and rf bypassing. The same is true of control shafts and shielding material that has oxidized.

## parasitics

This is a local problem and is usually confined to power amplifiers running in class B or C. Amateur handbooks explain the problem and a receiver sniffer can help to find the offending circuit.

Switching power supplies can generate RFI through "cross-current conduction." While this is mainly a problem for the supply,[10] it can react with other components to produce rf bursts at switching times. Cure the cross-current conduction and you generally cure the RFI.

Emitter followers sometimes have parasitic oscillations. A couple of hundred ohms in series with the base usually helps but proper design is better.[11,12] This type of spur results more in distortion of the signal; it is usually at vhf and rather unstable.

Spurious signals have many causes but can be identified with a little work. Most of them come from mixing so you can head them off while still in the block diagram stage. Others result when the schematic is converted to hardware. Careful circuit grouping and common sense will stop those.

Spurs are only for cowboys? Any experienced rider will tell you a good horse doesn't need spurs at all.

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**ham radio**

# high-dynamic range active double-balanced mixer

By Ulrich L. Rohde, DJ2LR

**During** recent years, much research has been done on solid-state mixer design. One important design configuration was described by Rafuse;[1] this circuit, using four single-gate mosfets, was incorporated into the Racal receivers built in the late 1960s. Even with the excellent intercept point of +28 to +30 dBm, this design suffered from the problem of excessive local-oscillator feedthrough back to the rf input. To cure the problem, an rf amplifier was added before the mixer. Another important contribution, in using field-effect transistors, came from Ed Oxner of Siliconix.[2] But even the eventual development of the VMP4 power mosfets did not solve all the problems of the device's high input and feedback capacitance.

As I explained in a previous article,[3] optimum performance of present-day active mixers can only be

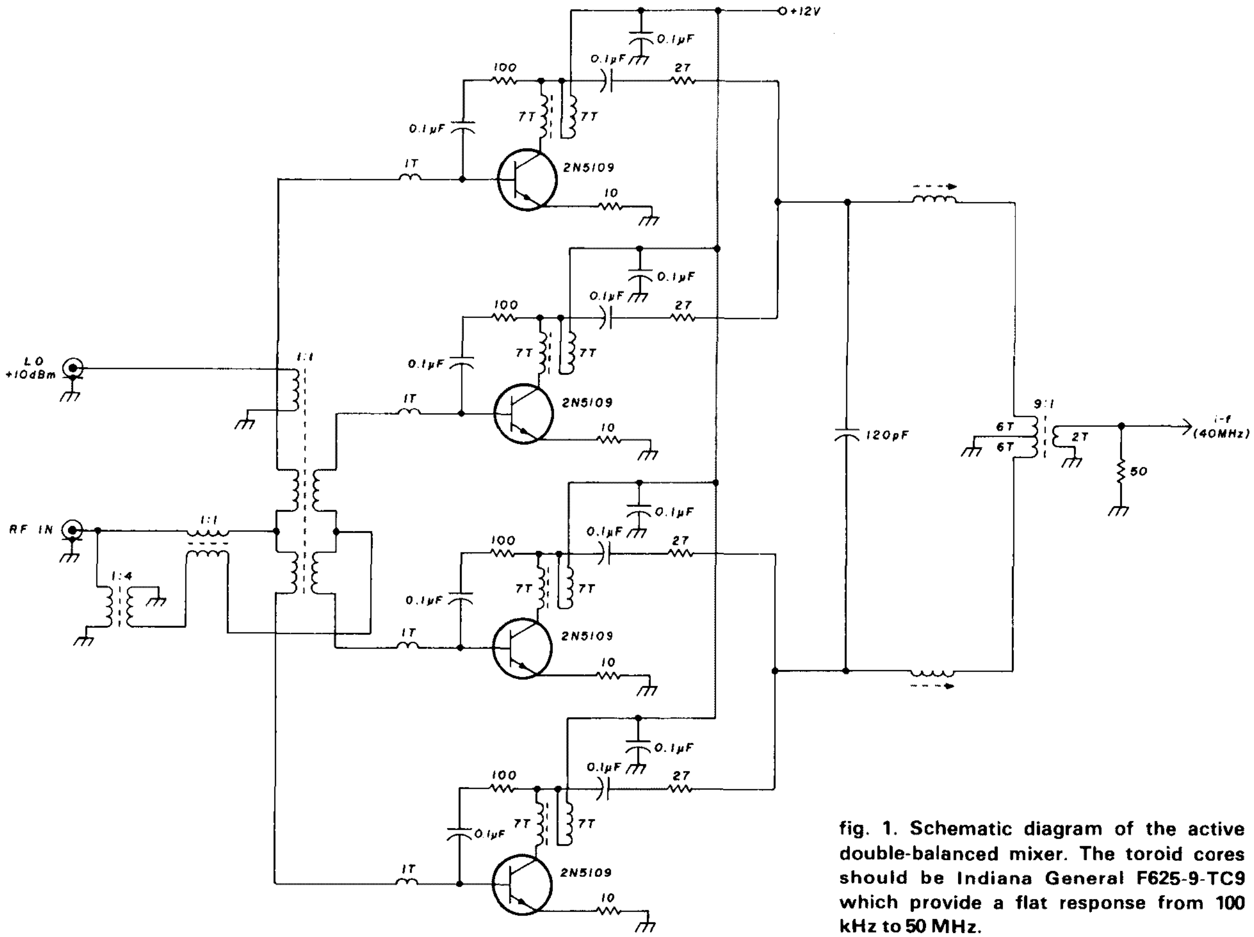
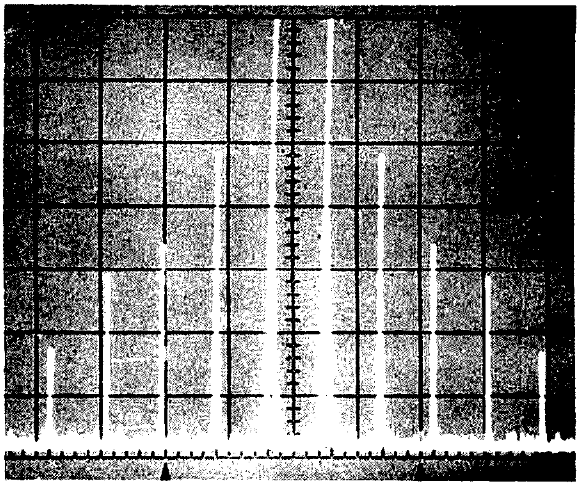
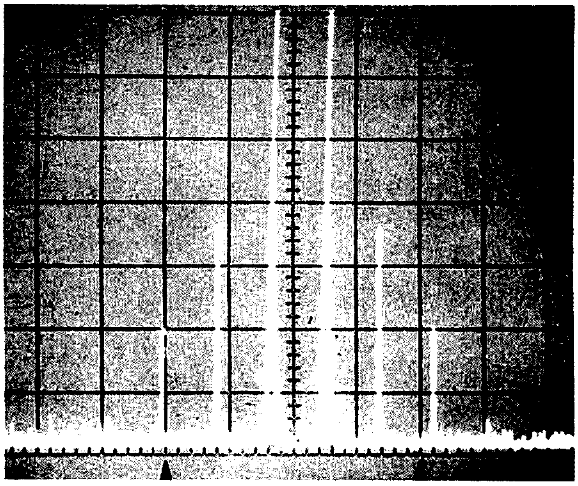
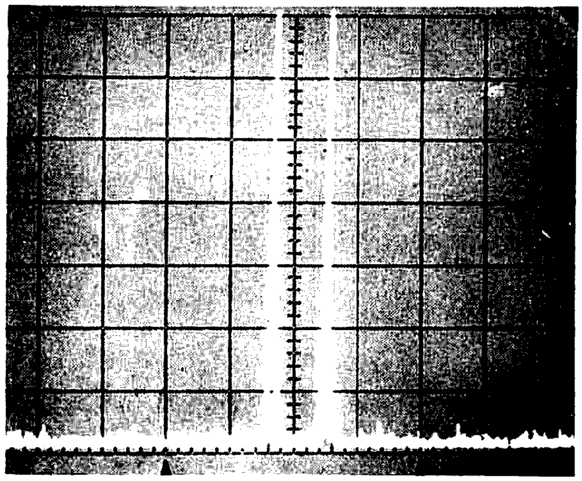



fig. 1. Schematic diagram of the active double-balanced mixer. The toroid cores should be Indiana General F625-9-TC9 which provide a flat response from 100 kHz to 50 MHz.







The three photographs show spectrum analyzer presentations of double-balanced mixers. In each case the input signals were 0 dBm. The top line in the graticule also represents 0 dBm. *A* shows an ordinary passive double-balanced mixer with +7 dBm LO drive. *B* is an active double-balanced mixer, made in accordance with the Siliconix applications note, using a U350 fet. The intercept point is +17.5 dBm. *C* shows the performance of the circuit in *fig. 1* run under the same conditions as *B*. In this case, the distortion products are suppressed 65 dB below the carrier levels. For the same LO-drive level, this means you have a 30 dB decrease in distortion products with roughly the same component costs and noise figure.

achieved when the i-f port is properly terminated. For example, if an i-f filter follows the mixer, the filter must present a constant impedance over a wide frequency range.[4] If not, the intercept point will deteriorate due to the high impedance levels, causing current or voltage saturation of the active devices. Therefore, the *passive* double-balanced mixer, together with a proper wideband termination using fets in a grounded-gate push-pull configuration, has yielded a superior performance over a wider frequency range.

Recent developments in the transistor field have produced a group of CATV transistors that are characterized by low noise figures and high gain-bandwidth products. By applying new types of rf feedback, linear operation can be obtained that will provide superior IMD product performance over previous fet and tube designs.

**Fig. 1** shows an active double-balanced mixer using four CATV transistors (2N5109) with a feedback circuit as described in reference 3. The rf feedback, together with impedance stabilization, avoids the drawbacks of the field-effect transistor mixer while giving a higher intercept point and stable gain. This increased performance is achieved with the same drive level as previous designs yet at the same time providing a noise figure of approximately 10 dB. The emitter resistors are used to reduce the amount of flicker noise in the system. With a local oscillator level of +13 dBm, a +40 dBm intercept point can be achieved. Lowering the LO level to +10 dBm causes little performance degradation but does decrease the distortion level.

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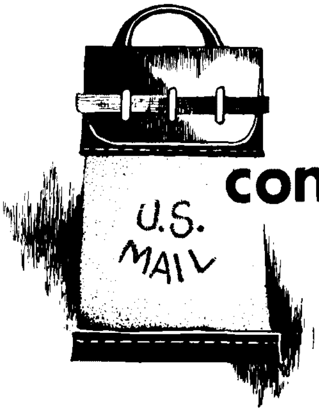



# comments

## silver plating

**Dear HR:**

Recently you published an item on electroplating copper wire with silver, which had the advantage of improving the appearance and freedom from corrosion (although silver tends to become silver sulfide), a varnish coating being applied to prevent the latter. It was also stated that, electrically speaking, the lower resistance of the silver was of little importance under 100 MHz. I have found the following:

**1.** Urethane varnish, diluted 1:6 with gasoline, makes a very durable coating so that silver plating may be unnecessary (for applications below 100 MHz).

**2.** Exhausted photographic fixer (fixer that requires at least twice the original time to clear the film [which can be made by fixing out waste film in a small amount of fixer, if desired]), will deposit a coating of silver upon clean, grease-free copper, by simply immersing it in the fixer. Used fixer can be obtained from firms doing *offset* printing; they use it for their negatives and usually dump it in the sewer when it becomes exhausted. Otherwise a friendly amateur photographer doing black-and-white work will be glad to donate his used hypo which is virtually worthless in small quantities.

To test photographic fixer for silver content, polish a piece of heavy copper wire or copper tubing with abrasive, then plunge it into the fixer — it will emerge with a bright coating if sufficient silver is present in the fixer solution.

**Lloyd Jennett, WA0AGD**
**Des Moines, Iowa**

## contact bounce eliminators

**Dear HR:**

The short article by W9KNI in August, 1976, *ham radio* (page 80) on how to eliminate contact bounce problems in keyers can be solved by a simpler method. There is no such thing as a switch that has no bounce, either on make or break. To overcome the problem Motorola introduced the MC14490 Hex Contact Bounce Eliminator about a year ago. Using this IC, up to 6 contacts can be debounced. The only other component required is a small capacitor. The IC will interface with CMOS or TTL and either normally open or normally closed contacts.

**Harry R. Hyder, W7IV**
**Scottsdale, Arizona**

## synthesizer design

**Dear HR:**

"Modern Design of Frequency Synthesizers" was one of the few articles, amateur or otherwise, that really got into the fine points of synthesizer design. The only point that I would take exception to regards the Motorola phase detector (MC4344). It is not a simple flip-flop, but a true frequency phase detector with infinite pull-in range like the CD4046. I have used both with essentially equivalent results.

The only difference is that the CD4046 has a charge pump. A better name would be tri-state output; the three states being raise frequency, lower frequency, and hold frequency. A flip-flop or shift register phase detector has only two states — raise or lower frequency. The CD4046 is capable of a full $V_{CC}$ to $V_{SS}$ output voltage swing. Unlike the MC4344 (output of $\pm 0.7$ volts) the CD4046 could be used to drive a VCO directly, without additional dc gain. The lack of external amplification could make a low-gain, one loop synthesizer, based on the CD4046, have a cleaner output.

**Jerry Pulice, WB2CPA**
**Staten Island, New York**

## internal resistance of Radio Shack meters

**Dear HR:**

In the July, 1976, issue of *ham radio* W0MAY comments on the lack of information about the internal resistance of Radio Shack panel meters.* This information was made available to the stores in the monthly technical newsletter that the company uses to keep employees up-to-date on the various product lines.

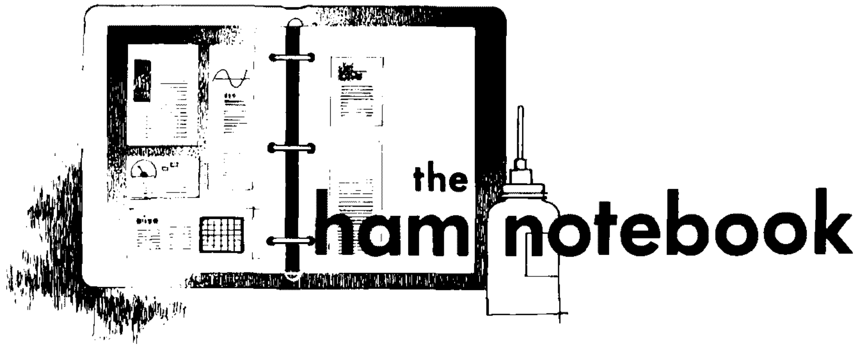
The list is reproduced below for the information of your readers. Several of the meters listed have been discontinued, but the information is included for those who may still have them lying around.

| catalog number | scale | internal resistance (ohms) |
|---|---|---|
| 22-016 | 0-150 Vac | 80 |
| 22-017 | 0-50 μA | 1600 |
| 22-018 | 0-1 mA | 80 |
| 22-019 | VU | 750 |
| 22-020 | "S" meter | 80 |
| 22-036 | 0-15 Vdc | 80 |
| 22-037 | 0-100 μA | 1000 |
| 22-051 | 0-50 μA | 1600 |
| 22-052 | 0-1 mA | 80 |
| 22-053 | VU | 750 |

It seems as though W0MAY's calculation of 1600 ohms for his 0-50 μA meter was right on the money! I hope your readers will find this list of use.

**G. L. Katzenberger**
**Troy, Ohio**

*David Cheney, W0MAY, "Shirt Pocket Transistor Tester," *ham radio*, July, 1976, page 40.




## simple crystal oscillator

National Semiconductor has a new IC designed to flash an LED from a single voltage cell. It is called LM3909N and is similar to the standard minidip. I've found that the LM3909 makes a very efficient crystal oscillator in the i-f range of 100-500 kHz. A 100 kHz crystal will generate strong harmonics beyond 30 MHz. Power drain is less than 0.5 mA at 1.2 volts; an AA cell should last for months.

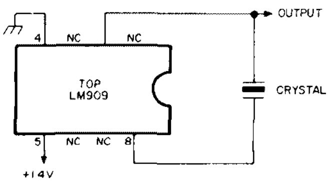



fig. 1. Schematic of the simple crystal oscillator using the National LM3909 integrated circuit. The crystal can be adjusted for its exact frequency by connecting a capacitor in series with pin 8.

All you need is the LM3909N, a minidip socket, a crystal, and a penlight cell. For ultra-low current drain, you may add up to 2000 ohms of resistance in series with the cell, which will drop the drain to about 0.25 mA.

For a low-power i-f signal generator, use a 465-kHz crystal. I couple the oscillator output to a receiver input via a 100-pF capacitor. For more precise frequency control, connect a capacitor in series with pin 8 of the IC. About 10 pF brought my 100 kHz crystal to zero-beat with WWV.

**Isaac Queen, W2OUX**

## fm-ing on uhf multimode transceivers

Some users of the popular synthesized vhf ssb/fm transceivers have occasionally noticed a frequency shift (fming) when the unit is operated on ssb. In my case, contact with the manufacturer of the transceiver failed to throw any light on the subject. The objectionable fming was very noticeable on voice and gave a peculiar wavering quality to the modulation. Inquiries among other local vhf operators showed that this fault was not unique, nor was it limited to a single brand of equipment.

A series of tests finally revealed that the fming was caused by a minute quantity of rf getting back into the frequency control circuits of the transceiver. Operation of the transceiver into a dummy load showed no fming, yet operation into the station antenna revealed the presence of fm when the transceiver was in the ssb mode.

The solution to the problem was two-fold. First, the swr on the transmission line from the transceiver to the antenna had to be reduced to a very low value in that portion of the band where ssb operation was used. Second, the line had to be brought away from the transmitting antenna field in such a fashion that no rf was induced into the outer conductor of the line. This meant relocating the line so that it dropped down directly beneath the transmitting antenna instead of coming away at an angle.

The problem of achieving a low swr across the entire two-meter band was solved by switching from a Yagi antenna to a log-periodic bandpass Yagi having an equivalent power gain. This antenna provided an swr value of less than 1.4:1 across the entire two-meter band. Using the log periodic Yagi, plus relocation of the transmission line, completely solved my vexing problem.

***Bill Orr, W6SAI***

## tower guying

Judging by some of the improper tower installations I've seen, it is clear that many amateurs do not understand the basic principles of tower guying. In addition to the very obvious function of preventing a tower from falling over in a high wind, guy wires also perform a second less obvious function: they prevent the tower from twisting, but only if properly installed. This is critical, because a tower with a large antenna load is vulnerable to twisting when the antenna starts whipping around in the wind. In most cases, a tower is much more likely to twist and buckle than it is to bend straight over.

Some people use nylon or polypropylene rope to guy their tower; they think that because it has the same ultimate breaking strength as the recommended steel guy wire, it's an acceptable substitute. The fallacy is that the rope will stretch and the steel guy wire will not. Also, the rope does very little to protect the tower due to twisting. Given a severe enough wind storm, it can come down with all guys intact.

Another common misapplication of

table 1. Data for determining tension in a guy wire.

| wire size | breaking strength | initial tension | mass slugs/foot | mass (kg/m) |
|---|---|---|---|---|
| 1/8 inch (3mm) HS | 1330 lbs. ( 603kg) | 150 lbs. ( 68kg) | $9.33\times10^{-4}$ | ($45.6\times10^{-3}$) |
| 3/16 inch (5mm) HS | 2850 lbs. (1293kg) | 300 lbs. (136kg) | $2.27\times10^{-3}$ | ($110.9\times10^{-3}$) |
| 3/16 inch (5mm) EHS | 3990 lbs. (1810kg) | 400 lbs. (181kg) | $2.27\times10^{-3}$ | ($110.9\times10^{-3}$) |
| 1/4 inch (6.5mm) HS | 4750 lbs. (2155kg) | 500 lbs. (227kg) | $3.73\times10^{-3}$ | ($182.3\times10^{-3}$) |
| 1/4 inch (6.5mm) EHS | 6650 lbs. (3016kg) | 700 lbs. (318kg) | $3.73\times10^{-3}$ | ($182.3\times10^{-3}$) |
| 5/16 inch (8mm) HS | 8000 lbs. (3628kg) | 800 lbs. (363kg) | $6.37\times10^{-3}$ | ($311.3\times10^{-3}$) |
| 5/16 inch (8mm) EHS | 11200 lbs. (5080kg) | 1200 lbs. (544kg) | $6.37\times10^{-3}$ | ($311.3\times10^{-3}$) |
| 3/8 inch (9.5mm) HS | 10800 lbs. (4899kg) | 1100 lbs. (499kg) | $8.49\times10^{-3}$ | ($414.9\times10^{-3}$) |
| 3/8 inch (9.5mm) EHS | 15400 lbs. (6985kg) | 1600 lbs. (726kg) | $8.49\times10^{-3}$ | ($414.9\times10^{-3}$) |

guying is to use steel guy wires but fail to tension them properly. Left too loose, even steel guy wire will not give adequate protection against twisting. The Rohn tower literature recommends tensioning each guy wire to 10 per cent of its ultimate strength. There is an easy method for determining the tension in a guy wire without resorting to the expensive dynometers.

A relationship exists between the tension in a guy wire and its resonant vibrating frequency. This relationship is defined by the formulas:

$$n = \frac{1}{2l}\sqrt{\frac{T}{m}} \quad \text{and} \quad T = 4n^2 l^2 m$$

where

- $T$ is the tension in foot pounds,
- $n$ is the vibrating frequency in Hertz,
- $l$ is the length of the guy wire in feet, and
- $m$ is the mass per unit length of the guy wire in slugs per foot (1 slug = 32.17 pounds).

To find the resonant frequency, start the wire swinging back and forth, count the number of full cycles in ten seconds, and divide by ten. It will probably be necessary to push on the wire with each swing to keep it going for this length of time. The data in **table 1** for different sizes of seven-strand galvanized steel guy wire is from the Rohn catalog. Using this data and the appropriate formula, the actual tension in any guy wire, and the vibrating frequency for the recommended tension can easily be determined.

There will always be cases where there is insufficient room to place guy anchor points as far out from the base of the tower as the manufacturer recommends. While an installation such as this is always something of a structural compromise, there are other means to help overcome this deficiency. The easiest technique is to use torque arms. These are assemblies that attach to the tower at each guy point, and the guy wires are in turn attached to the torque arms. For guying purposes, these make the tower more resistant to twisting. Another method is to use four guys at each level, spaced every 90 degrees instead of the usual three with 120 degree spacing. Other possibilities would include guying at more levels than the minimum recommended, using heavier guy wire, and making the guy anchor points stronger.

Put up your tower as high as you want, but guy it right. Then when the ice storms come and the hurricanes blow, you can relax. Your antenna may blow away, but a properly guyed tower will stay up through just about anything.

**John Becker, K9MM**

## SB200 CW modification

Improved CW operation of the Heath SB200 amplifier can be obtained by reducing the key-up plate current to 0 (class C operation instead of AB). This simple modification will maintain the same amplifier output, and substantially increase tube life by reducing key-up idling power from about 200 to zero watts. The modification is performed by adding a 250-ohm, 10-watt resistor in series with the ANTENNA RELAY lead, (**fig. 1**). By adding the resistor, the grid bias voltage, which is normally developed across a 33-ohm resistor,

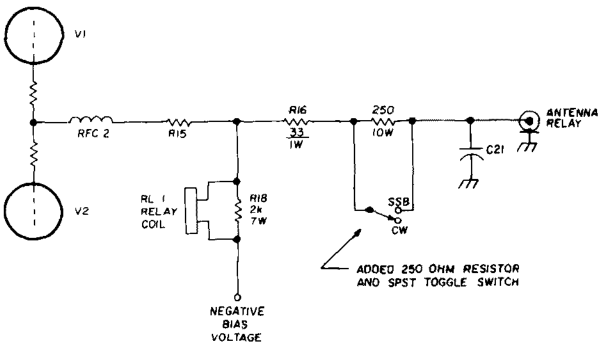



fig. 2. The tubes in the amplifier are held cut-off by the addition of the 250-ohm resistor since bias is developed across the resistor. A toggle switch can be used to short out the resistor for ssb work. If this is not done, excessive distortion will result because the amplifier is operated in class C.

is increased. Fortunately, the resulting reduced current is still adequate to operate the antenna relay, which is in series with the negative grid bias supply output.

When ssb operation is desired, the 250-ohm resistor is simply shorted out. The resistor can be mounted internally, with a small spst toggle switch mounted on the front panel between the meter selector switch and RELATIVE POWER SENSITIVITY control shown in **fig.** 2.

**John Abbott, K6YB**

## 32S-series ALC meter improvement

The drifting zero adjustment of the Collins S-Line ALC meter is quite common and annoying. It's due in part to the components involved, but basically this syndrome is caused by a change in line voltage which produces a corresponding change in the reference voltage taken from the screen of V3, the i-f amplifier. This voltage is in turn divided down by the combination of R22 and R23 and precise zero balance is accomplished by R20, the ALC meter zero adjustment potentiometer. Therefore, as the low-voltage supply changes, the zero adjustment also varies. When negative, it's most disturbing.

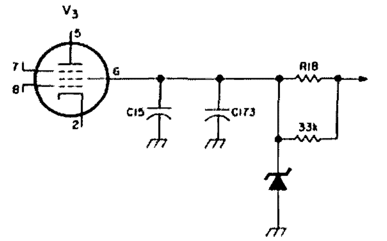



**fig. 3. The addition of the zener diode helps stabilize the screen voltage on the i-f amplifier in the Collins S-line, which improves ALC metering. The diode is a 100 volt, 500 mW zener.**

The solution is a rather simple change which also has a secondary beneficial effect. The change is shown in **fig.** 3, with the two additional components being shown with the heavy lines. First, a 500-mW, 100-volt zener diode (IN5271) is added from pin 6 of V3 to ground, and a 33k, 1-watt resistor in parallel with the existing screen dropping resistor, R18. The screen voltage is now reasonably well regulated, providing a relatively stable reference voltage for the ALC meter circuit. Also, because the developed ALC bias voltage is impressed on the grid of V3, the screen voltage rises as the plate current is reduced during normal peak speech excursions. The result is that the transconductance of V3 tends to rise as the ALC bias voltage simultaneously seeks to reduce the gain of the tube. With the added regulation, this potential problem is eliminated. The results have been quite favorable in my Collins 32S-3 with the ALC zero varying no more than plus or minus one minor division during five months of operation.

**Marv Gonsior, W6FR**

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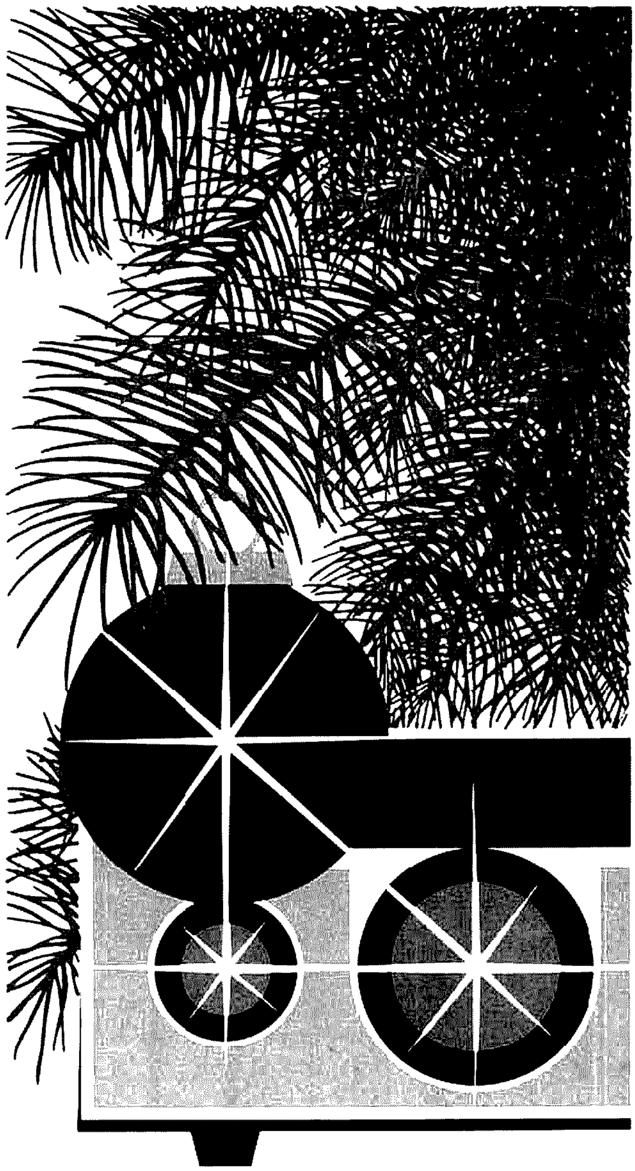
# ham radio

magazine

## DECEMBER 1977





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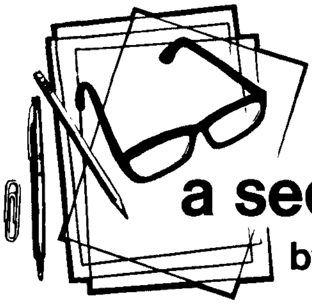


# a second look

by Jim Fisk

**During the holiday season** it's customary to take stock, to look back over the past year, and to make our resolutions for the next — resolutions, no doubt, which will be forgotten by the time the snow melts from the landscape and the trees begin to show their buds. The long winter nights are also a good time to plan that new antenna system or to dream about some new station equipment. With the snow swirling up to the window sills and the cold winds howling down from the north, perhaps it's a good idea to take some time to think about where amateur radio has been, and where it's going.

With the World Administrative Radio Conference (WARC) of 1979 now less than two years away, I can't help wondering what our amateur bands will look like in the 1980s. Will amateurs be given some of the additional high-frequency bands requested by the WARC planning committees, will the width of the amateur bands be pared down, or will we lose much or all of our high-frequency allocations? Nobody will know the answer to that until the final votes are tallied in 1979, but I suspect it will fall somewhere between the two extremes.

There are some who would have you believe there will be *no* high-frequency amateur bands after 1980, and very little vhf spectrum either, but I'm more optimistic than that. Optimism, unfortunately, leads to apathy and that, my friends, is our worst enemy. Perhaps it's best to prepare for the worst and approach WARC '79 with cautious optimism.

It must be remembered that the last international conference which had much effect on the high-frequency spectrum was held in 1947 when the United States and our Allies had considerable influence on the 50 member countries of the United Nations. Radio amateurs were highly regarded by our government for the part they played in war-time communications — not as amateurs, but because they provided a pool of trained technicians and communicators. To a lesser extent the same thing was true in Britain and the Soviet Union. Radio Amateurs were also the backbone of the communication networks set up by the resistance movement in Europe, and of the coast watchers in the South Pacific.

Governments which had severely curtailed amateur radio before the war now recognized its great potential as a national resource. Amateur radio was no longer considered a nuisance to be tolerated, but an activity which should be encouraged. Part of that encouragement was a new, exclusive 15-meter band. Old timers will hasten to point out that bits were shaved off the top ends of 10 and 20 meters, and 160 meters was dominated by Loran, but most amateurs agreed that 15 meters more than made up for the losses.

By the time the next ITU conference on high-frequency allocations was convened in 1959, the United States' sphere of influence had decreased and it looked like amateur radio was in serious trouble; the foreign broadcasters wanted big chunks of 40 and 80 as well as portions of 20 and it was uncertain if we could rally enough votes to save amateur radio. Fortunately some of the nations who weren't particularly friendly toward the United States but supported amateur radio came to the rescue, with the result that the amateur bands in the Western Hemisphere came through unscathed (amateurs in other parts of the world lost 50 kHz of shared space on 40 meters).

In general, the United States and other governments which were supportive of amateur radio in 1959 still are, but in the 20 years since that last conference the balance of power has changed; the emerging nations are now in the majority and they are not altogether in favor of amateur radio — a few ban it outright. Many of these nations have few amateurs, so to them the amateur bands represent wasted space — space they feel should be allocated to a radio service that better serves their national interest. These are the same countries which often oppose the policies of the rich western nations simply because it's in the vogue to do so.

Nevertheless, there's still hope, because many of the questions to be asked at WARC '79 will be answered on the basis of their scientific merits. There's bound to be a certain amount of political arm twisting, but if the delegates from the emerging nations can be made to see the value of an amateur radio service to the technological development of their nation, perhaps they can be persuaded to vote in favor of increased amateur spectrum.

**Jim Fisk, W1HR**
**editor-in-chief**

# presstop

ARRL'S "CODE OF ETHICS" has been challenged in a formal written complaint filed with the Federal Trade Commission's Bureau of Competition.

Specific Complaints are that the Code will violate anti-trust laws by restraining trade, constitutes a "deceptive practice" as defined by the FTC, violates the First and Fourteenth Amendments to the Constitution, and that vendors who sign the ARRL pledge will become accessories after the fact in the above violations.

FCC'S DROPPING OF DOCKET 19759, the proposal that the 220-MHz Amateur band provide a home for a new CB service, doesn't mean that the band won't still become the new CB home. It does remove the immediate threat to the band, however, and many opinions have it that the longer the decision on where CB should go is delayed, the less likely 220 becomes as a choice.

In Announcing Its Termination of Docket 19759 the Commission pointed out that so many changes in related circumstances have occurred since several thousand comments were filed on it back in 1973, those comments were now obsolete. However, the question of a new CB band and where to put it is still very much alive, and 220 will undoubtedly be one of the options when the Commission considers the issue again in a future rulemaking.

A PETITION FOR RECONSIDERATION of the FCC's Report and Order on repeater deregulation (Docket 21033, Presstop, November), is being prepared by the ARRL. In it three issues will be emphasized: restoration of the WR-prefixed callsigns for repeaters, restoration of repeater licenses, and the need for formal consideration of the needs of the so-called "weak signal" vhf/uhf operations.

Plenty Of Support for the League position appears likely, as many repeater groups already oppose the dropping of repeater callsigns and licenses. In addition, reservations over the repeater sub-band expansion and even the proposed new bandplan for 144.5 - 144.5 MHz is starting to build among FM users as well as various SWOT and other VHF/UHF user groups.

AMATEUR LICENSING IRREGULARITIES will receive a full-fledged investigation run by an FCC Administrative Law Judge. The decision to go all-out on such a probe was reached at a closed meeting of the Commission, when information was presented that some Amateurs had apparently paid for the issuance or upgrading of their licenses or for special callsigns; that some of the same abuses may have occurred without payment; and that some Amateur callsigns have been issued inconsistent with normal FCC procedures.

ARTHUR C. CLARK, the noted science-fiction writer, was made an honorary AMSAT member in ceremonies attended by most of the AMSAT brass — Clark's honor came in recognition of his predictions of communications satellites and synchronous satellites in a 1945 Wireless World article.

1978 Orbital Prediction Booklets for OSCAR 7 only will be available shortly from Skip Reymann, W6PAJ, Box 374, San Dimas, California 91773. They're free to AMSAT Life members who request them; $3 to AMSAT Annual members and $5 to non-members — be sure to include AMSAT membership number and an self-addressed label with orders.

OSCAR 7's Mode Schedule will be changed effective January 1 to two days in Mode B for every day in Mode A, and the new schedule will be shown in W6PAJ's orbital calendar coming out in December. Ample Mode A operations will be provided by the Russian's "RS" spacecraft and AO-D, and OSCAR 7 is considerably more sensitive in Mode B than it is in Mode A.

A New Satellite Bandplan is also going into effect January 1 which will place CW only on the bottom third of the satellite downlink, mixed CW/SSB in the center third, and SSB only operation on the top third — the reverse of current practice.

OSCAR 6's Fifth Birthday was October 15, but revival efforts from VE3SAT failed to bring any response from it. RIP.

The Amateur Space Program made the Congressional Record in October when K7UGA lauded it during a Senate discussion of Sputnik's 20th anniversary.

THOSE PACIFIC AND CARIBBEAN prefix changes may not be as drastic as originally announced. FCC's news release announcing the change has now been "cancelled," and while it appears that prefix changes will still be made they'll be done in such a way that the resulting callsigns should identify the individual islands or island groups (Presstop, November).

FCC'S "GAG" ON DISCUSSIONS of current matters will remain in place as a result of the Supreme Court's decision not to review the Court of Appeals decision in the "Home Box Office" case (Presstop, June).

WESTINGHOUSE SCIENCE TALENT SEARCH is open to any high school student in the United States and Puerto Rico who'll graduate before October 1, 1978. Teachers who have an outstanding student who'd qualify for one of the many scholarships and awards must request entry materials from Science Service, 1719 N. Street, Washington, D.C. 20036 — entries are due by December 17, 1977.

# present-day receivers

## — some problems and cures

Some thoughts on and cures for problems encountered in modern amateur communications receivers

**The modern-day communications** receiver is going through a continuous evolution that has brought about significant improvement in certain operating features. Among these are greatly improved frequency stability and setability, better selectivity, a slow and consistent tuning rate from band to band, and a wide-range automatic gain control system that functions on CW and single sideband. At the same time, unfortunately, the design philosophies which have made the above advances possible have also reduced the typical receiver's ability to simultaneously handle weak desired and strong undesired signals. This absolute reduction in receiver dynamic range has occured at the same time the number of high-power signals on the amateur bands has been increasing.

Insufficient dynamic range in a receiver can result in one or more stages being over-driven into nonlinearity by undesired strong signals. The result is internally-generated intermodulation distortion (IMD) products. These undesired products can occur in any mode of operation, but are easiest to identify on CW. Two CW signals which are overdriving a receiver will generate IMD products, but only when both stations are transmitting simultaneously. In the extreme situation, not only may IMD occur, but one signal alone can block, deaden, or desensitize the receiver.

In a pileup or contest situation, many strong CW stations can cause serious receiver overload, intermodulating with each other, and resulting in multiple phantom signals; it will appear as if several operators are randomly tapping their keys, or that you are listening to the Novice band with a diode detector without a BFO.

Two or more ssb signals with the correct frequency relationship can also intermodulate with each other and result in IMD products on top of the station you are listening to. The interference, however, will be unintelligible. IMD can also occur from a single ssb station on an adjacent channel as the individual speech frequencies mix with their own harmonics. Generally speaking, transmitted IMD from an rf power amplifier will be worse than that internally generated in the receiver, with the result that the transmitted IMD may cover up a receiver's shortcomings. An operator may never be certain whether the unintelligible signals he hears are being generated within his receiver, or coming from the outside — there is enough rf interference to contend with without the receiver creating its own!

The improvements mentioned in the first paragraph have been generally obtained by using a double- or triple-conversion scheme, plus a non-bandswitched master oscillator (PTO or VFO). Depending on the design technique, the first i-f may have a bandwidth of as much as 500 kHz, as in the Heath SB-104, or as narrow as 6 kHz in the Drake R-4B. Assuming that most of a receiver's selectivity occurs at the second intermediate frequency, you might think that the wider the bandwidth of the first i-f, the greater the chance of picking up more strong signals which could overload the second mixer. Of greater importance than this bandwidth, however, is the *net gain* between the antenna and the mixer that drives the narrow crystal or mechanical filter.

The Collins R-390A, for example, has three mixers and two separate gain stages ahead of its mechanical

**By J. Robert Sherwood, WB0JGP, and George B. Heidelman, K8RRH,** Sherwood Engineering, Incorporated, 1268 South Ogden Street, Denver, Colorado 80210

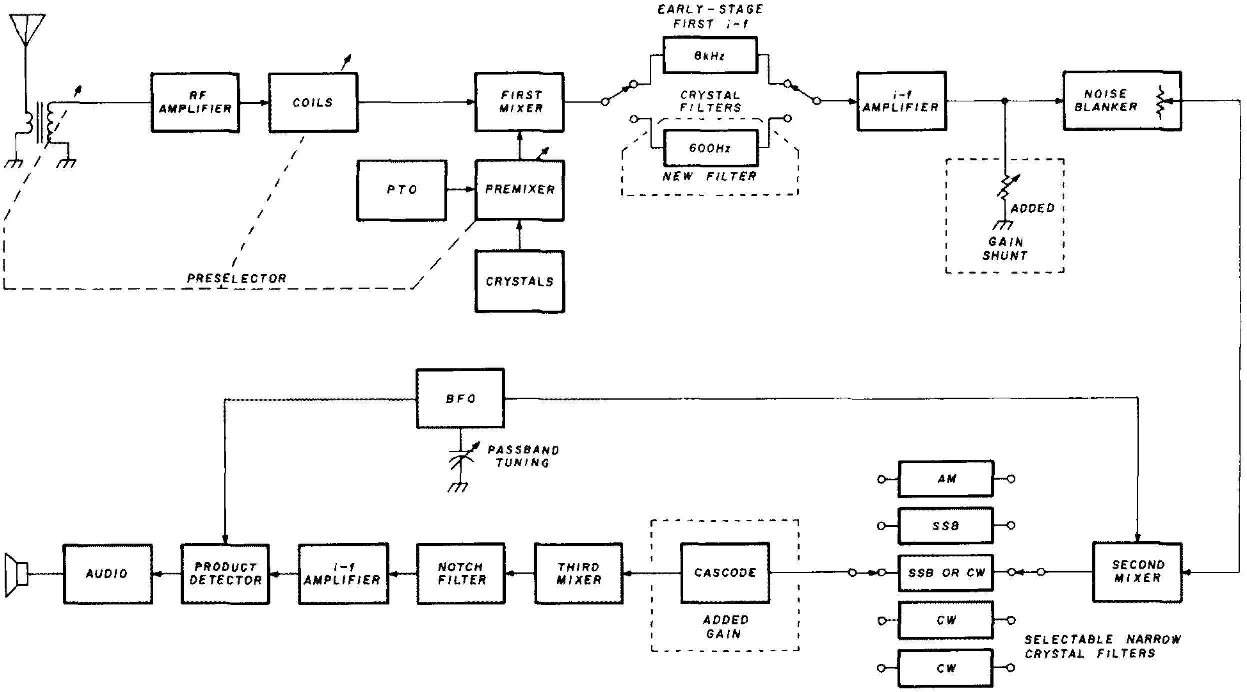



**fig. 1. Block diagram of the Drake R4C receiver showing the gain redistribution. A shunt across the first i-f amplifier will reduce its gain the same amount as is added after the narrow i-f filters.**

filters; it also has a set of elaborate, mechanically-tracked tuned circuits which have high $Q$ and high insertion loss. Thus the net gain from the antenna to the major selectivity-determining elements is low enough to maintain good dynamic range.

Another receiver, the Heath SB-303, has a 500-kHz wide first i-f *window*, but unlike the R-390A, it has little selectivity ahead of its narrow filters and too much gain. This results in higher susceptibility to overload from strong signals anywhere in the band, which then cause undesired IMD products to be generated within the receiver.

At the opposite end of the bandwidth scale is the Drake R-4C with its 8-kHz wide first i-f filter at 5645 kHz. This four-pole crystal filter does an excellent job of keeping most of the undesired signals in the band from passing on to a second high-gain mixer. However, any undesired strong signals that *do* pass through this 8-kHz *window* can proceed to the second mixer with disasterous results. The net gain from the antenna to the narrow second i-f crystal filter can be as high as 50 dB when a desired weak signal (S1) is being received; this puts an impossible demand on the i-f stages, since the 1-dB compression point of the second mixer output has occured with any signal 30 dB over S9. An undesired signal, outside the narrow selectivity but inside the first i-f *window*, that is S9+40 dB (−33 dBm or 5 mV across the 50-ohm antenna input) for example, would have to be linearly amplified to a level of +17 dBm (1.58 volts across the 50-ohm narrow-filter input) and then be rejected by the filter. To supply this power level to the filter, the high-impedance plate of the second mixer would have to linearly swing more than 40 volts to yield a signal that is as great as 15

One topic that has received considerable attention by amateurs in recent years has been that of receiver performance and design. Many approaches have been covered, from the initial design of the "super receiver" to modification of existing equipment; but to the person with just a casual interest, the reasons behind some designs may not be readily apparent. In fact, the problems themselves may not be noticeable to the ordinary amateur. This article is another in a continuing series that shows you how to recognize the problems in typical modern receivers; in addition, it discusses modifications applied to one receiver and the motives behind these changes.

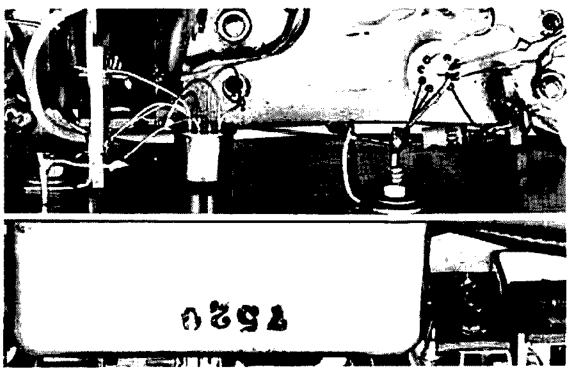
Of major importance is the reason for the modification. The intent of this article *is not* to prove that one particular receiver is superior to another for whimsical reasons, but to realistically and fairly compare different receivers by presenting test results on comparable circuits. On the basis of the test results, design changes were made in one receiver in an attempt to improve overall performance. You will notice while reading the article that the results are given in very specific terms; this will help you to better understand the basics of receiver performance standards. With this knowledge, *you* will be able to judge the merits of the different receivers on the market and choose one according to your own needs. **Editor**

volts rms; even if this level could be produced in a low noise mixer, which is highly unlikely, the filter could be damaged.

What actually results when there are two undesired signals at S9 + 40 dB with the correct frequency relationship, over loading the second mixer, is a spurious third-order IMD signal that is greater than S9 in strength. This would certainly be strong enough to obliterate the desired weak signal!

One possible reason why such net-gain design errors are overlooked is our present method of testing receiver dynamic range. This subject has received considerable attention lately in *ham radio*[1,2,3] and *QST*.[4] An increasingly popular method of testing for dynamic range has been developed by Wes Hayward, W7ZOI, and is used by the ARRL.[5] Basically, it consists of applying two *well-isolated*, equal-strength signals, 20-kHz apart, to a receiver's input and then adjusting their level so that the undesired third-order IMD products generated within the receiver are just equal to the noise floor of the receiver. The difference in level between the noise floor and the test signals gives the receiver's dynamic range. The higher the receiver's dynamic range, the better it can handle both desired weak and undesired strong signals at the same time.

The choice of 20-kHz spacing for the two test signals is arbitrary and in many cases satisfactory. In a receiver which has all its significant selectivity far down the i-f chain, this signal spacing is relatively unimportant. If the early-stage bandwidth is narrower than the test signal spacing, however, its selectivity will partially or completely reject one or both of the test signals, resulting in a highly inflated dynamic range reading. We feel these measurements should cover worst-case conditions since real-life interference on the amateur bands may be spaced less than 20 kHz.



**Installation of the 600-Hertz first i-f filter. The filter is installed on a vertical shield near the original 8-kHz filter. The devices with 8 leads are TO-5 size relays that are used to select the appropriate filter.**

Third-order IMD products, with 20-kHz spacing, will occur 20 kHz below the low frequency test signal and 20 kHz above the high frequency test signal. When the receiver is tuned to a third-order internally-generated spurious IMD signal, the test signals are 20 and 40 kHz up or down the band. The 25-kHz-wide crystal filter in the first i-f of the Signal-One transceiver, to name just one example, will greatly attenuate the test signals before they can reach the following stages. Thus, 20-kHz spacing will test only the front end and first mixer. What is needed is spacing narrow enough so that both test signals can pass through any selectivity prior to the narrow filter. We feel a spacing of 2 kHz will satisfy this requirement, and at the same time be wide enough so the narrow filter will adequately reject the test signals when the receiver is tuned to an IMD product.*

The Drake R-4C, with its 8-kHz-wide first i-f filter, shows an inflated 20-kHz dynamic range of 83 dB. This reading has remained quite consistent over several receivers, including one we tested at the ARRL laboratory.† When the test signals are placed 2 kHz apart, however, so they *both* pass through the 8-kHz filter, the dynamic range drops to around 58 dB.

## improving receiver performance

There are three ways to improve a receiver's dynamic range. If the second mixer cannot handle the required level, one option is to replace it with a mixer that will do the job. Unfortunately, as WB4ZNV discovered,[6] the process of replacing an active mixer with the superior passive double-balanced mixer is a laborious task, even if it does improve the receiver's overload characteristics. Oscillator injection levels and impedances are usually not compatible with existing circuitry.

Another remedy is to redistribute the gain in the receiver, reducing it ahead of the overloaded stage and building it up again after the narrow filter. A third method is to insert more early-stage selectivity into the receiver so strong interfering signals are not as likely to get past the first mixer. We chose to inves-

*When performing a 2-kHz IMD test, one very important factor must be taken into consideration: the noise sidebands of the signal generators. General test equipment, oscillators, or VFOs are more than adequate for testing, until a receiver's dynamic range nears 100 dB. At this point it will be impossible to accurately measure true receiver IMD products if the signal generators are producing excessive low-level spurs and noise. At this time there are only two or three generators that have the necessary sideband suppression; one manufactured by Hewlett-Packard and another by Rohde and Schwartz.

†The ARRL laboratory uses a pair of AN/URM-25 signal generators to perform IMD tests. A 2-kHz IMD test produced results within 2 dB of those obtained by the authors while using the high quality, low-noise sideband Rohde and Schwarz XUA signal generator.

tigate the latter two options, using our own R-4Cs.

The initial gain redistribution began with a 20-dB reduction of the signal level as seen by the second mixer. This gain loss was then restored after the narrow filters at the high-impedance grid of the third mixer. The original amplifier used a single jfet plus a step-up transformer to provide the necessary gain, but the circuit suffered from instability problems and noise. It was then decided to relocate the added gain outboard from the receiver and insert it at a convenient 50-ohm point, the output of the switchable second i-f crystal filters (see **fig. 1**).

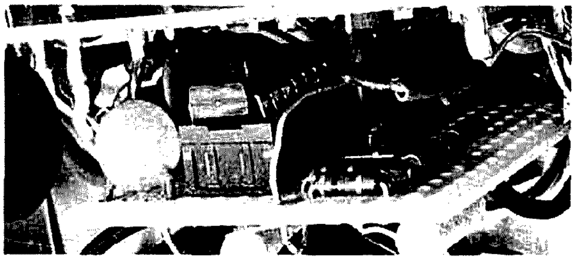
A cascode jfet amplifier, with 50-ohm input and output impedances (**fig. 2**), was built and inserted into the i-f chain just prior to T-6. The coax cable that connects T-6 and the mode switch was lifted at the switch end; two lengths of miniature coax (RG-174/U) were then run out through a slot in the rear of the receiver. The first length is connected to the lugs on the mode-switch wafer, while the second is spliced into the cable that feeds the transformer.

This amplifier can possibly be located inside the receiver. Regardless of its location, it should be mounted in a metal box or other well-shielded enclosure. Two toroidal transformers provide the necessary impedance changes, their associated trimmer capacitors forming resonant circuits. While both trimmers can simply be peaked for maximum signal, the input may be fine-tuned for the best compromise signal-to-noise ratio among the switchable narrow filters. (The 2N5950 and 2N5953 jfets may be purchased from G. R. Whitehouse Company, Amherst, New Hampshire 03031).

We found the best way to attenuate the signal level into the second mixer was to swamp the output of the first i-f amplifier Q1 (V3/6BZ6 in early receivers). A miniature 5000-ohm multi-turn trimmer, from noise blanker socket pin 4 to ground, made a convenient way to adjust this level. Simply adjust the trimmer to drop the calibrator signal 20 dB on the S-meter; then adjust the gain pot on the cascode amplifier to restore the S-meter to its previous level. On certain receivers it may be necessary to peak T-6 to obtain 20 dB of gain from the cascode amplifier; always readjust both cascode trimmers after making a gain change.

If the noise blanker is installed in the receiver, significant IMD products can occur in its stages, too. Due to noise limitations, however, the blanker cannot be starved a full 20 dB. Instead, after replacing blanker resistor R1 with a 0.001 $\mu$F disc capacitor, reduce the gain to the blanker about 12 dB, and then turn down the blanker output pot 8 dB to achieve the 20 dB reduction at the second mixer. Alternately, the gain of blanker transistor Q2 can be decreased by reducing its emitter resistor bypass capacitor, rather than readjusting the blanker output pot.

Take care not to use too much cascode amplifier or blanker gain; otherwise amplified 5645-kHz oscillator leakage can degrade system performance. With the antenna disconnected and the top and bottom covers of the receiver in place, make sure the S-meter does not kick upward more than one-quarter S-unit when the passband tuning is slowly turned through its range. In some receivers it may be necessary to jumper the cable-braid ground point of the Q4 oscillator board with a short clip lead to the shield tray on which the blanker board rests to reduce this oscillator leakage to an acceptable level. It might also be necessary to insulate the frame of the rear carrier-oscillator jack from the chassis ground.



**The new product detector is installed next to the audio transformer and behind the variable capacitor used for passband tuning. The entire assembly is mounted on a 1-3/4 x 1-5/8 inch (4.5x4.1cm) board.**

Also, if the cascode amplifier breaks into oscillation when the mode switch is between detent positions, reverse the leads of a high impedance winding of one of the toroids.

Proper operation of the gain redistribution circuits provided greatly reduced susceptibility to IMD overload problems on both CW and ssb, as was visibly demonstrated with strong nearby DX contest signals; yet the receiver was still able to meet its sensitivity specification. Agc attack distortion was also reduced somewhat. Dynamic range improved from 58 dB to around 70 dB, while using our 2-kHz spacing test method.

## i-f filters

As an additional CW remedy we chose to increase the selectivity (possibly on a switchable basis) following the output of the first mixer; the bandwidth is presently determined by an 8-kHz wide four-pole crystal filter. This bandwidth is needed on phone to pass an upper and/or lower sideband signal. A bandwidth of at least this magnitude is also required to pass undistorted noise pulses to the blanker. A noise blanker's usefulness, however, is marginal at best with one or more strong nearby signals, due to its agc greatly increasing the blanking threshold, or possible false triggering. Thus, the need for narrowing first i-f selectivity ahead of the noise blanker,

which reduces blanker effectiveness, occurs under conditions which are usually unfavorable to blanking in the first place.

Circumstances could occur where blanking would be necessary at all times, such as when you suffer from a continuous very high level of blankable noise. In these cases, the 8-kHz first i-f filter must remain ahead of the blanker. Then a properly-terminated narrow filter could be inserted just *after* the blanker, but before the second mixer. The signal path can be switched between the narrow filter and an attenuator equal to its loss. While the chance of second mixer overload is greatly reduced with this arrangement, there is no such narrow bandwidth IMD protection for the blanker; this limits the receiver's potential dynamic range considerably below what is otherwise obtainable. It is therefore mandatory to use the cascode gain redistribution system with this special, optional filter arrangement. With this arrangement close-in dynamic range will be in the high 70s.

We decided that the first i-f CW selectivity should be equal to the widest desirable under contest conditions. We then designed a new 600-Hz six-pole filter, keeping in mind package size limitations and insertion loss requirements. We've also developed a miniature relay system which allows instant interchange of our internally-mounted, CW-bandwidth, first i-f filter with the existing 8-kHz phone unit.

The project of minimizing overload in the R-4C was now complete and totally successful. When measured using our worst-case 2-kHz test method, the receiver's dynamic range jumped from an original unacceptable 58 dB to a final excellent 85 dB. This value ranks with the best of the commercially-available amateur gear on the market today, and

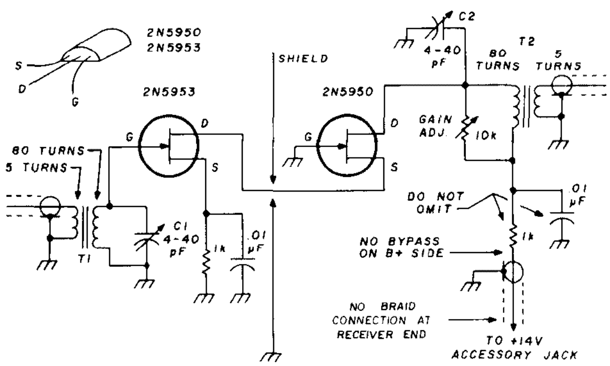



**fig. 2. Schematic diagram of the cascode amplifier used for the gain redistribution. There is only one ground return on the circuit, through the input coax cable. The braid on the output coax cable goes to the primary of T6 which is not grounded at that point. T1 and T2 are wound on Micrometals T-50-2 toroidal cores. The high-impedance windings are 80 turns of no. 30 AWG (0.25mm) while the low-impedance windings are 5 turns of no. 24 AWG (0.5mm).**

should be more than adequate for most practical situations. As a side note, a similar arrangement of first i-f filter switching can be used on ssb by inserting a set of 2.6 or 2.3-kHz phone filters in the first i-f for improved phone selectivity.

## simple receiver testing

While we made use of a considerable amount of test equipment during this project to measure dynamic range, you can make comparative tests using only a crystal calibrator and transmitter vfo, *loosely* coupled into the receiver. Comparative noise floor measurements, with no antenna connected, can be made by measuring the preselector noise peak (above later stage noise) with an ac voltmeter connected to the audio output line.

When making gain redistribution or selectivity changes, adjust the receiver to maintain its original net gain by measuring the calibrator level on some specific frequency. We use 7.2 MHz as our reference frequency. Here the calibrator level should read about 15 to 20 dB over S9 with nothing connected to the antenna input. (Don't readjust the S-meter sensitivity pot.) Two strong test signals, accurately set to a specific S-meter level, will produce a repeatable reference IMD that can also be measured on the S-meter. As improvements are made the IMD, read on the S-meter, will drop. We made our 2-kHz tests at S9 + 40 dB, and ended up reducing the IMD from greater than S9 to less than S3.

## filter rejection

The 600-Hz first i-f filter, in addition to greatly reducing the chance of overload, had the extra benefit of eliminating the annoying signal leakage around the narrow second i-f filters. This problem of not being able to realize the ultimate rejection capabilities of a well-designed filter is one that plagues all equipment that, to our knowledge, is presently on the market. It is really quite difficult to even design a test fixture to correctly measure the ultimate rejection of a filter. Obtaining adequate ultimate attenuation, which should be in excess of 100 dB for an eight-pole filter in a receiver or transceiver, requires tedious attention to detail. Current ground loops and stray capacitive coupling are the main problems that must be eliminated. We have had many frustrated amateurs ask us to provide a filter for their receiver or transceiver which would not leak like the factory installed units. Unfortunately, some of the limitations were in the receiver and not the filter. Although replacing or adding to an existing *late* narrow filter can often considerably improve skirt selectivity, the only way to eliminate the last traces of these leakage problems, in existing popular receivers, is to add a filter earlier in the set with a

bandwidth closer to that of the main filter. The early filter should preferably be on a different frequency from the later one, such as in the R-4C or 2B.

We tested one all-solid-state American transceiver that had so much leakage around the CW filter that a 2-kHz dynamic range test could barely be made. The IMD was masked by the test signal leakage until special audio filtering was employed.

While discussing filters, we would like to emphasize the importance of a great variety of bandwidths being available to the operator. Most of the equipment on the market has just one standard phone bandwidth, with one CW filter available as an option, and when installed it must be used at all times for that mode. Many of the imported rigs are examples of these limitations. The Yaesu FT-101B has only a six-pole 600-Hz filter, and the Kenwood TS-820 is limited to only a six-pole 500-Hz unit.

By today's standards a six-pole 500-Hz filter is quite broad and has a poor shape factor. One possible reason for offering only these filters is that the design of the equipment was based on the use of an ssb filter having an insertion loss of only 2 to 4 dB. Unless a manufacturer employs special technology in building, say, an eight-pole 350-Hz filter that is more advanced than required for a phone filter, the insertion loss will rise to an unacceptable 14 to 16 dB. It is quite undesirable to have the signal drop 12 dB when the CW filter is used; a compromise is made, and the six-pole filters mentioned above are offered. Even with this trade-off, there is an additional insertion loss of 5 to 7 dB compared to the phone filter, and relatively poor skirt selectivity.

As a minimum, the receiver net gain should be designed around the lossiest filter, with the losses of the other filters increased to that constant level. Another school of thought suggests that the noise integrated by each of the filters should be the same, requiring increasing gain (or decreasing insertion loss) as narrower filters are selected. To our knowledge, no amateur equipment manufacturer is currently keeping the integrated noise constant, and only the R-4C provides for constant insertion loss with narrow bandwidth filters.

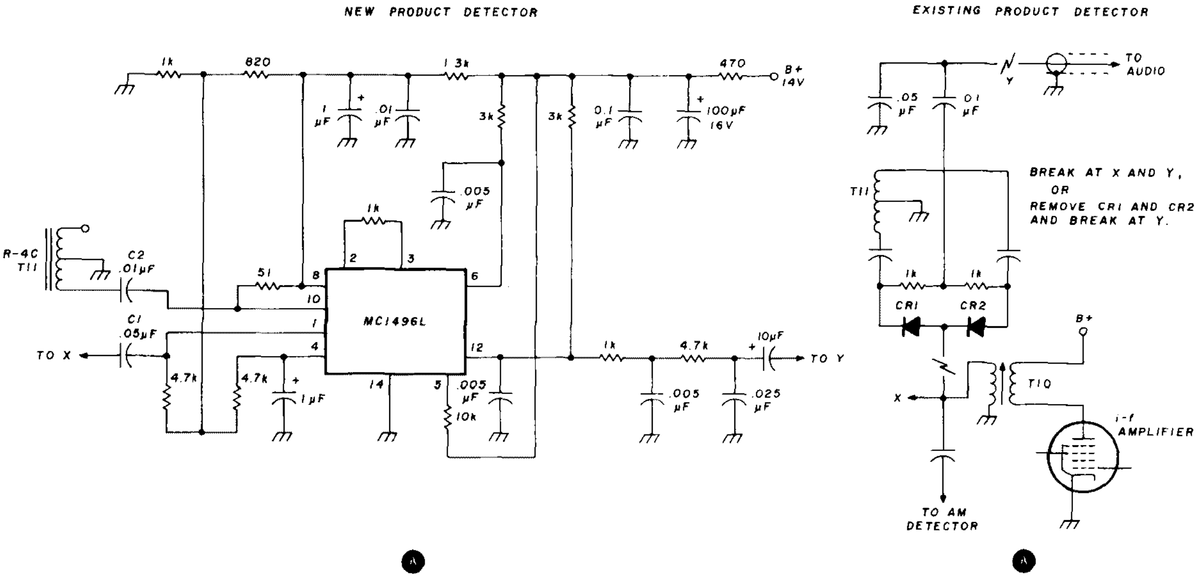
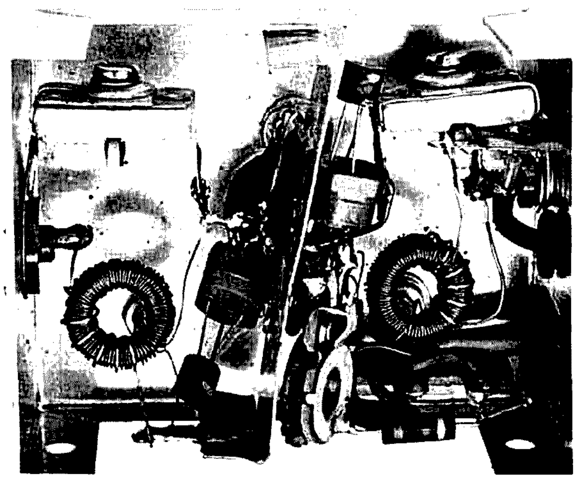



**fig. 3. The MC1496L can be used as a product detector as shown in A. The IC plus associated components are mounted on a small circuit board which is installed next to the audio transformer in the receiver. C1 and C2 are critical values and should not be substituted. For smaller size, the 1-μF capacitors may be tantalum. B shows the interconnections between the detector and the receiver.**

We have noted with interest the comments from some of our Japanese and German filter customers about American rigs such as the R-4C and T-4XC. The cost of these units in their home countries, due to import duties, is 30 to 50 per cent higher than here in the United States, but the discriminating foreign amateur is willing to pay that premium partly because of the excellent filters which are available. Compared with the typical filter in the average set, the Drake eight-pole 250-Hz and the Sherwood eight-pole 125-Hz CW filters are valuable assets. Similarly, an optional 1500 to 1800-Hz ssb filter* can make the dif-

*Drake also offers the FL1500, a 1500-Hz filter. Though publicized as an RTTY filter, it provides exceptional performance, especially under difficult phone contest conditions. **Editor.**



**Cascode amplifier used for gain redistribution is installed in a small enclosure. The shield must be in place between the stages of the amplifier.**

ference in being able to hold a contact under heavy interference and contest conditions.

It takes some practice to become proficient at using a narrow i-f filter, just as in learning to tune with the wide-skirted audio filters. But during crowded band conditions a 250-Hz filter can often be too broad! One CW operator used the 125-Hz filter in his R-4C almost exclusively during the hectic 160-meter contests.

The entire line of filters for the R-4C is excellent and can be adapted to any receiver or transceiver. A construction article in the 1977 ARRL *Handbook*[7] describes a method of adding bandpass tuning to a receiver lacking this feature. This circuit uses 455-kHz filters and is inserted in the receiver i-f chain by converting down to 455 kHz and back up again. This basic idea can be used with any pair of filter and receiver intermediate frequencies.

You could convert from 3395 kHz up to 5695 kHz and back down again, for example, or down from 9 and up again. As the difference between the two i-f frequencies becomes smaller, the difficulty of the conversion process increases. A Drake R-4B owner who wishes to add R-4C filters to his receiver has to cope with a conversion frequency difference of only 50 kHz. Howard Sartori, W5DA, has developed a circuit for use in his R-4B which can be adapted to any i-f by simply changing one crystal oscillator. It has been used on intermediate frequencies as low as 50 kHz and as high as 30 MHz with excellent results. His circuit is described on page 20 of this issue of *ham radio*. One precaution, when adapting the *Handbook* circuit or W5DA's i-f converter to a transceiver: make sure the transmitted signal does not have to pass through the added filters. Otherwise, with use of the two narrowest filters (the FL-250 and CF-125/8), the transmitter carrier offset frequency adjustment would become quite critical, and keying on the transmitted signal could be too soft.

The Kenwood TS-820, which we have in the lab, has a noise floor and dynamic range in the ssb mode that is virtually identical to that of the Drake R-4C. Both units perform very well on phone; when you want to dig out a weak CW signal on a quiet band, however, the R-4C is significantly better. The R-4C's gain remains constant when a CW filter is switched in, but the TS-820's drops off 5 to 6 dB. Even if a weak received signal is above the noise floor, this gain reduction increases the agc threshold to the point where it may become necessary to manually ride the gain control. The Yaesu FT-101B we tested had a dynamic range, at any test signal spacing, as bad as the unmodified R-4C when measured with the *worst-case* 2-kHz test method. The bulk of the problems in the FT-101B were caused by a bipolar transistor in the noise blanker which was being overdriven.

A receiver's maximum net gain from the antenna to the detector can change significantly from band to band without having much effect on the measured sensitivity. Two sets with similar signal requirements for a given signal-to-noise ratio can have vastly different capabilities in handling weak, fluctuating signals, especially on the 10- and 15-meter bands. As the net gain falls off, more and more signals will fall below the agc threshold. The R-4C, for instance, holds a much more consistent net gain from 80 to 10 meters than the TR-4C. The TS-820 increases the net gain on 10 meters compared to 20 and 15 by changing a capacitive tap on the rf amplifier drain. Its gain, however, is too high on 160 meters, resulting in a higher susceptibility to overload by broadcast stations. When connected to a nearly self-resonant 160-meter vertical antenna at our lab in Denver, the TS-820 grossly overloads with the eighteen local broadcast stations, developing more than 1 volt across its antenna input. Without the 20-dB rf attenuator switched in, the 160-meter band is nothing but a solid mass of S9 + 30 dB IMD products.

The TS-820's front end is not selective enough to cope with this admittedly unusual receiving situation. On 1.8 MHz, the preselector attenuates signals that are 100 kHz off frequency by 18 dB. In comparison, the R-4C attenuates these same signals by 38 dB. On 3.6 MHz, the TS-820's front end is down 8 dB at 100 kHz off frequency, the TR-4C by 12 dB, and the R-4C by 24 dB. When tested on 10 meters, the *500*-kHz attenuation is 8 dB on the TS-820, 8 dB on the TR-4C, and 15 dB on the R-4C.

One way to eliminate the need for a sharp preselector is to use an up-conversion scheme, with the first i-f above 40 MHz. The input may only need a

bandpass filter that rejects signals below 1.8 and above 30 MHz. Then image signals would fall above 80 MHz and be virtually eliminated by the bandpass filter. The first mixer must have a much greater signal-handling capability than in present receivers, however, because it would see all stations between 1.8 and 30 MHz. Two strong local signals, one on 14 and the other on 21 MHz, could produce a 7-MHz IMD product.

The R-4C and the TS-820 show a 20-kHz test-signal-spacing dynamic range in the ssb mode of about 80 dB when tested on 20 meters. At this frequency, the preselectors do not significantly enter into the dynamic range test, since they will not attenuate the test signals more than 1 dB. This is not the case on 160 meters, especially with the R-4C. Here, its high-*Q* front end attenuates the 20-kHz signals enough to raise the dynamic range by 12 dB. On the other hand, some receivers have too much gain on 80 and 160 meters which, even with sharp preselectors, could yield a dynamic range no better (or even worse) than on 20 meters.

While the 20-kHz dynamic range of the R-4C improves on the lower frequencies because of its preselector, the 2-kHz dynamic range measurement remains quite constant at just under 60 dB. Similarly, it is consistently above 83 dB with the 600-Hz first i-f filter that cures its *window* overload problem. The TS-820 does not have this *window* problem since it is a single-conversion design and has no overloadable stages between the wide noise blanker filter and its narrow filter. Any improvement in dynamic range with increasing frequency separation of the test signals can only be attributed to its preselector.

A detailed review of the TS-820 in *CQ-DL*,[8] far more comprehensive than anything published in this country, showed a 6-dB improvement in dynamic range as the test signal spacing was increased from 2 to 50 kHz. It is interesting to note that *CQ-DL* also feels that a close-in 2-kHz spacing is necessary for proper evaluation.*

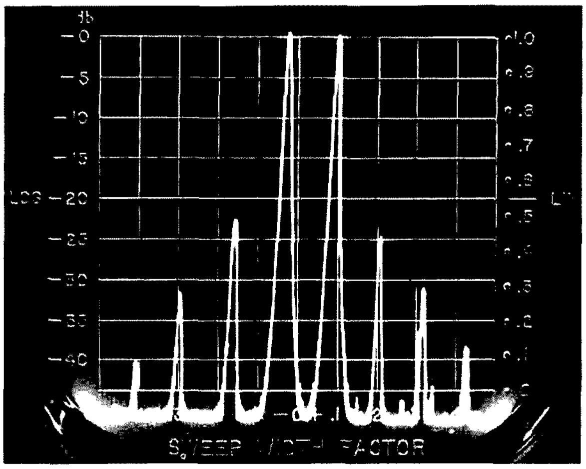
The Atlas 210X, without its noise blanker operational, has a better than average dynamic range of about 90 dB, which would be even better if its double-balanced mixer were properly terminated above the i-f frequency.[2] This could be accomplished with the use of a diplexer, as described by Wes Hayward,[4] or with a power jfet, as related by Ulrich Rohde.[2,3] There is one limitation in the 210X that cannot be easily remedied, however; its potential strong-signal handling capabilities cannot be fully realized due to its noisy conversion oscillator. Since this oscillator has noise sidebands that are only 65 dB down 10 kHz on each side of its center frequency, all the signals passing through the mixer will take on similar noise sidebands. Consider a strong station near a desired signal that is weaker in amplitude. Reciprocal mixing of oscillator noise can cause noise sidebands to be transferred to the strong nearby station and cause interference to the desired signal. Thus, even if the i-f filter's ultimate rejection is actually realized in the receiver circuitry, which is doubtful in practice, this high level of rejection can be negated by wide-band mixer noise. So while it takes two strong signals to cause IMD which can interfere with weak signal reception, a noisy oscillator and one strong signal can cause the same unfortunate results.[9]

The noise blanker in the Atlas 210X also degrades its dynamic range, diminishing the advantage of the double-balanced passive mixer. The 210X transceivers we tested had a dynamic range of between 73 and 81 dB, depending on the band selected. When the blanker was turned on, these numbers dropped by 3 dB.

There is little reason for a noise blanker to include additional gain stages which can degrade receiver performance. The TS-820 has only a 4-diode balanced blanker gate in its i-f chain; therefore, it does not reduce the overload capability or significantly increase the noise floor. Alternately, a balanced mixer or push-pull i-f stage can be gated for noise blanking; this requires no additional gain stages in the signal path.

## product detectors

Another area that could use additional work is that of the product detector. As the name implies, its output should be the product of the two input signals. If



**IMD generated at the output of the R-4C second mixer by two 5 mV signals at the antenna input. The signal spacing was 2 kHz. The receiver was tuned so that the narrow second i-f filter was positioned away from any test signals or IMD products. Therefore, with no signal reaching the AGC, the receiver gain is at maximum and the S meter reads S1.**

*A recent independent measurement by DJ2LR showed the intercept point of the TS820 to be − 12 dBm.

BFO injection is removed, output should go to zero. If this is not the case, as in the Heath HW series, envelope detection is also occurring, which causes audio distortion. On the other hand, the 6GX6 product detector in the Drake R-4, TR-4, and TR-4C, and the 6BE6 in the Drake 2A and 2B, works very well.

Other extraneous outputs can occur even if the detector is acting solely as a product mixer. A detector should be a double-balanced, or other arrangement, which provides good isolation between input and output. The two-diode detector in the R-4B and R-4C is not a double-balanced design and allows the detected audio to leak back and envelope modulate the last i-f stage. This resultant signal is detected in the agc, which then tries to follow it at an audio rate, especially (but not only) when the faster time constants are in use. This audio output sounds slightly distorted, and is noticeable on ssb as well as CW. In addition, BFO injection is marginal, causing additional distortion on AGC attack.

We decided to replace the product detectors in our R-4C receivers, but wanted to use a device that was compatible with the existing drive and impedance levels. The MC1496L active double-balanced mixer looked like a good choice, and with minor circuit changes from the data sheet, was installed in the receiver. The modulation of the i-f by the detected audio was eliminated, resulting in cleaner sounding audio. AGC attack distortion was further reduced.

The MC1496's main drawback is its high number of associated components. Eleven 1/4-watt resistors, nine capacitors, and the IC had to be squeezed on a 1-3/4 by 1-5/8 inch (4.5x4.1cm) board which was nestled between the audio output transformer and the adjacent PC board (see **fig. 3**). All R-4C owners, whether they change product detectors or not, should add a 0.0015 $\mu$F capacitor across R83 in the audio amplifier. This corrects a phase error in the feedback circuit, and eliminates an undesirable peak in the audio frequency response which accentuates harmonic distortion. The Kenwood TS-820 and the Atlas 210X both use a double-balanced diode product detector that works quite well, and needs considerably fewer parts, but they are low-impedance devices not easily adapted to some circuitry.

## conclusions

We have discussed several popular receivers and noted some of their strengths and weaknesses. Some problems can be corrected in the field, while others go beyond the scope of a weekend project. We've also investigated two ways to improve a receiver's susceptibility to overload, so that it can better handle today's high-level rf environment: redistributing the gain and increasing the early-stage selectivity with an additional filter. The importance of having a wide choice of adequate narrow filter selectivity, without leakage, was also mentioned. While most of our circuit changes have been applied to one specific popular receiver, the Drake R-4C, the ideas can be extended to other sets. A method of checking a receiver's overload capabilities which requires no test equipment was also described. Thus receiver changes can be evaluated as to their effect on dynamic range.

The real key to how a receiver performs is its net gain distribution, particularly in relation to the location of selectivity determining elements. A receiver must have a great deal of gain from its antenna to the speaker to be able to receive weak signals. But if too much gain is placed ahead of a narrow filter, the receiver is bound to overload and generate interference of its own.

How a receiver will perform in real-life situations can be determined in the lab, but only if it is tested in a manner that approximates the real world. We feel that the present 20-kHz signal-spacing method can be quite misleading, and should be augmented with our 2-kHz test procedure. If the two readings are significantly different, then further investigation *is* warranted.

As we stated at the beginning of this article, receivers have improved in many ways, especially over the past 15 years; at the same time, dynamic range has diminished. Amateur radio operators should be demanding excellence in this critical parameter. Improvements in receiver versatility need not reduce system performance, as we have so often observed. Potential problems can be eliminated in new equipment by state-of-the-art design or by retrofitting existing receivers. All that will be lost is some internally-generated rf interference!

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**ham radio**

# an up/down filter converter — matches any bandpass filter to any receiver i-f

Design and construction of an up/down converter that will interface any crystal filter with any receiver i-f. A design example shows how to add a 125-Hz filter to the Drake R-4B

**Ultimate receiver performance** is viewed by most amateurs as a moving target, with increasing cost just one factor that keeps the target out of reach. New inventions and techniques are constantly being announced by the fast-moving electronics industry; yesterday's dream of an ideal receiver becomes history long before the final receiver payment is due. Giant strides in IC technology have made receivers comfortable and easy to use through the addition of synthesizers, diode switches, and frequency counters. However, most *real* receiver performance improvements, in terms of signal-handling capability and selectivity, are still to be made. The name of the game is picking the weak signal out of the interference caused by many nearby strong signals. Then receiver performance specifications such as third-order intercept point, dynamic range, receiver desensitization, and mixer overload suddenly come to mind.

One goal of modern high-frequency receiver design is to process the desired signal through the narrowest available filter, with the smallest number of active components. Maintaining a credible noise figure, however, tends to legislate against throwing out all of the active front-end components except the mixer.* While giant strides have been made in semiconductor development, filter technology has been advancing rapidly, too.

This article will discuss the use of available crystal filters and will show you how to easily add high-performance filters to receivers without facing the frustrations of mixer design — frequency conversion, loss of sensitivity, and degradation of dynamic range.

## filter characteristics

A complete line of filters optimized at the same center frequency for CW, RTTY, ssb, and a-m, particularly the i-f in your receiver, is hard to find at a price you can afford, especially from a single manufacturer. Many receivers place the ultimate selectivity (that filter which passes the information bandwidth, such as a 2.4 kHz ssb filter) in the second i-f stage, or further down the active component chain than is desirable.

Frequently a receiver manufacturer does not offer filters which are optimized for RTTY or CW. If you find a filter with the desired response characteristics, chances are that it won't match the receiver i-f.

If you look to filter manufacturers who specialize in only crystal filters, you'll find that, within the past year or two, excellent crystal filters have become available that will optimize filtering for any mode of radio communications. The first stumbling block is the wide variety of filter center frequencies, typically

*Recent developments in solid-state mixer design actually permit the omission of all active stages prior to the mixer. It is now possible to obtain a mixer noise figure of 10 dB or less on the high-frequency bands. This will be sufficient for all but the most demanding reception requirements, such as OSCAR 7, Mode A on 10 meters. **Editor**

**By Howard Sartori, W5DA,** 721 James Drive, Richardson, Texas 75080

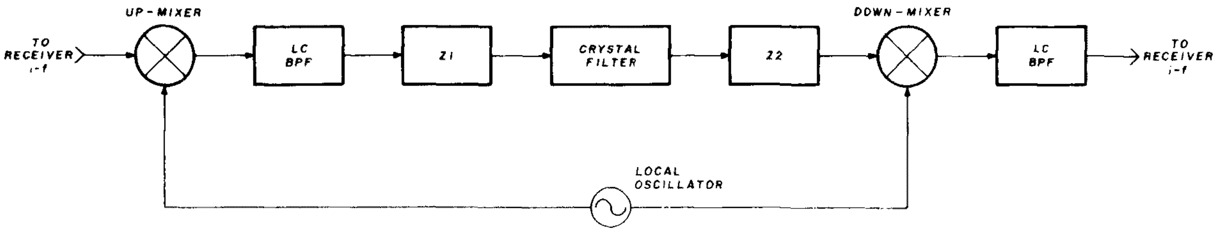



**fig. 1. Block diagram of the up/down frequency converter for adding an additional filter to a receiver. Z1 and Z2 represent the impedance matching networks necessary to interface the mixers and filter.**

in the range from 5 to 11 MHz. Rarely does one manufacturer produce a complete line of filters, optimized for the information bandwidth of each of the operating modes used by amateurs.

Cost has been a major factor in the past, but crystal filter production techniques have vastly improved and costs have turned downward. To put cost into perspective, and to consider the effects of inflation, excellent 8-pole crystal filters, with signal rejection *floors* below 100 dB, are now available for about the same price level as 4- and 6-pole crystal filters were about 5-10 years ago.

Describing a crystal filter by its shape factor, normally defined as the ratio of the 6-dB to 60-dB bandwidths, doesn't tell the whole story. This measure of squareness is typically 1.7 to 2.2 for a good quality ssb filter, the slope of the response curve in a simple filter is determined by the characteristics of the crystals. For a 2.4 kHz ssb filter with a shape factor of 1.75, for example, the attenuation/Δ frequency of the filter slope would be 54 dB/900 Hz. When narrower filters were designed, the bandwidth between corners was reduced but the slope remained essentially the same; a 500 Hz filter had a shape factor greater than 4.0. In fact, the bandwidth of the slope itself on one side of the ssb filter was wider than the 60 dB bandwidth of an optimized CW filter!

Eventually new fabrication techniques, such as mounting all crystals on the same header, permitted development of high-performance crystal filters. In some filters the entire passband may move as much as 2 Hz/°F, but this is not objectionable when the entire filter shape moves. To further illustrate the tremendous achievements in crystal filter technology that have occurred during the past several years, consider the 125-Hz CW filter now on the market. At one time, not too many years ago, 125-Hz crystal filters were a novelty of the laboratory. Today CW filters with 125-Hz bandwidths and shape factors of 2.5 are available for approximately $125.

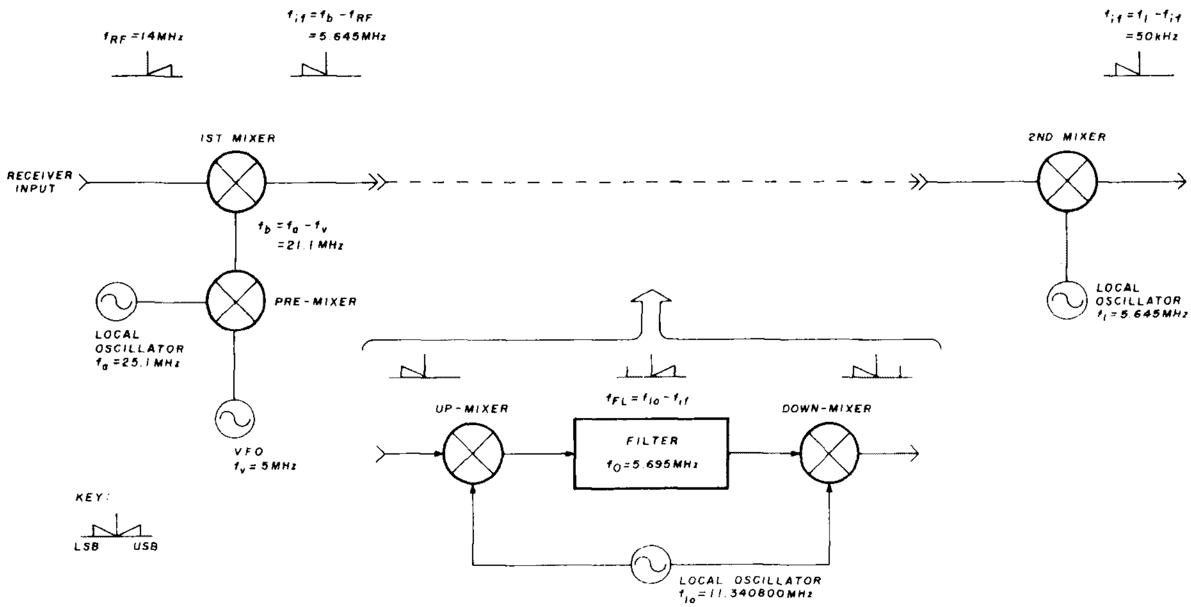



**fig. 2. Block diagram of the Drake R-4B showing the receiver frequency conversion scheme. The conversion is necessary for the additional filter inserted between the first and second mixer. This diagram can be used to check for any sideband inversion.**

By using the newer 8-pole crystal filters, receiver performance can be imoroved in the following ways:

1. Filters are optimized for the information bandwidth.

2. Signal-to-noise ratios are improved.

3. Filters placed as close to the front end as possible improve dynamic range.

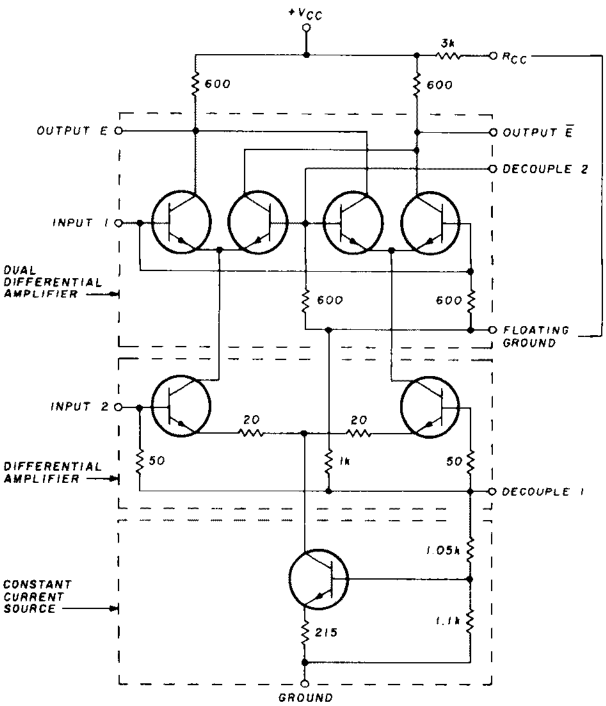



**fig. 3. A representative schematic diagram of the Texas Instruments TL442 doubly balanced mixer IC. Input 1 has an input impedance of 600 ohms, while input 2 is 50 ohms. The output can be connected for 600 or 1200 ohms output impedance, depending whether the supply voltage is connected to $+V_{cc}$ or OUTPUT E.**

4. Cascading with an existing filter to achieve better filter skirt and out-of-band performance.

When you optimize filters, you not only reduce susceptibililty to interference — you also reduce listener fatigue. Improved signal-to-noise ratios are particularly noticeable on the low bands for several reasons. First, if only an ssb filter is available for CW, when a 125-Hz bandwidth filter is switched into the circuit, the bandwidth improvement ratio will be 10 log 2400/125 or 12.8 dB. Even if a 500-Hz CW filter were available, a 125-Hz filter would improve the signal-to-noise ratio by 6 dB. In addition, by the very nature of impulse and static noise, the filter will prevent overloading the following stages; this means that the signal-handling ability of the receiver has been improved.

Some receivers have a nominal 2- or 4-pole filter in the first i-f, and the final selectivity in the second or last i-f stage. Obtaining all the needed selectivity at one i-f is difficult, however, because of signal radiation and leakage (even with the best shielding). By placing a high-performance crystal filter in the first i-f, overload of the second mixer can be greatly reduced. Further, spreading the selectivity over several stages is an excellent way to improve ultimate signal rejection.

Since the crystal filter you want to use will probably not agree with your receiver i-f, to say nothing of the input and output impedances, a convenient method is required to interface additional filters. Many receivers have a simple general purpose filter in the first i-f with an output impedance of 500 to 1000 ohms. This is an ideal place to add an outboard crystal filter.

Designing a conversion scheme can be a complicated process; consider all the variables. **Fig. 1** shows a block diagram of the general approach for heterodyning the first i-f signal up or down to a high-performance crystal filter, and then heterodyning back to the receiver. The first mixer (up converter) can easily overload; the down-mixer is not nearly as susceptible to overload. In addition, the local oscillator can act as a source of spurious radiation for birdies in the amateur bands. Since only one local oscillator is generally used for the two mixers, it may serve as the leakage path for the signal around the filter. Or, the local oscillator signal could feed through the down-mixer into the next i-f stage. And consider the fact that since almost all commercial crystal filters are in the 5 to 11 MHz frequency range, the receiver i-f and the crystal filter frequency should be very close.

Having both the receiver i-f, $f_{if}$, and the filter, $f_{FL}$, at nearly the same frequency is probably the leading cause for abandoning the project. Mixers not only mix; they can amplify. Therefore, if $f_{if}$ cannot be filtered out by the LC bandpass filter (BPF) at the output of the up-mixer, overload may eventually become a problem because both signals could be substantially amplified. To make matters worse, the conversion process usually results in some loss of desired signal; as much as 30 or 40 dB difference between the two mixer output signals, $f_{if}$ and $f_{FL}$, is not uncommon. As a result, the filter signal rejection floor is greatly diminished. If the two frequencies are within several hundred kilohertz, the isolation of $f_{FL}$ by the bandpass filter may be as big a task as manufacturing the high-performance crystal filter in the first place.

Finally, the filter must be very carefully matched to the up-mixer output and the down-mixer input. These matching networks are shown in **fig. 1** as Z1 and Z2.

Selecting a frequency conversion scheme should be done with care. **Fig. 2** shows the first i-f signal, filter i-f, and local oscillator frequencies. It is well to consider the particular sideband, too, since sideband reversal may not be desirable. **Fig. 2A** shows the complete receiver conversion scheme, with the Drake R-4B used as an example. The sideband slope diagrams show the relative sideband with respect to the incoming rf signal.

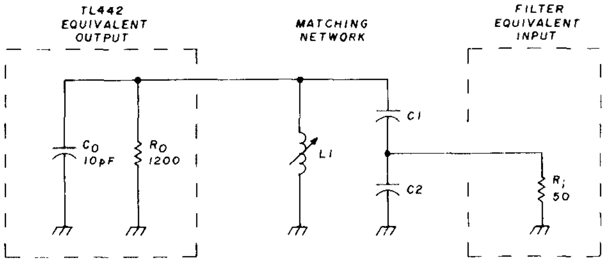
**Fig. 2B** shows how the up/down filter converter is integrated into the receiver's first i-f. Using the filter's center frequency, the required local oscillator frequency can be determined from $f_{LO} = f_{FL} \pm f_{if}$. The Sherwood Engineering CF-125/8* CW filter center frequency is 5695.0 kHz. Assuming an 800-Hz tone, the incoming frequency, $f_{if}$, would be 5645.0+0.8=5645.8 kHz; the local oscillator frequency would be 5695.0+5645.8=11340.8 kHz (or 5695.0−5645.8=49.2 kHz). The 11340.8 kHz frequency is, fortunately, not in any amateur band; but the 49.2-kHz local oscillator signal would fall within the passband of the receiver's second i-f! Therefore, 11340.8 kHz will be used as the local-oscillator frequency.

## solving the problems

The close proximity of the receiver's first i-f and the crystal filter frequencies was the toughest problem to solve. Convenience and easy-to-do were words which guided this design project for more than six months. Building the LC filter at the output of the up-mixer, shown in **fig. 1**, however, was anything but easy. Combining it with the impedance matching network, Z1, was complicated and certainly not repeatable without diligent tuning. Doubly balanced mixers were considered, but many of them required large numbers of external components and even null adjustments.

Finally, a doubly balanced mixer IC was found that provides internal preset nulls in excess of 30 dB, for

*Sherwood Engineering, 1268 South Ogden Street, Denver, Colorado 80210.




$$Q_u > \frac{R_S}{2\pi f L} \qquad C_T = \frac{C1\,C2}{C1+C2} + (C_O)$$

$$R_i = \frac{R_O}{\left(\frac{C2}{C1}+1\right)^2}\ \text{OHMS} \qquad Q_L = \frac{R_S}{2\pi f L} \qquad L_1 = \frac{1}{(2\pi f)^2 C_T}$$

**fig. 4. The capacitive tap-down network is used to match the up-mixer to the crystal filter. It can also be used to match a high-impedance first i-f stage to the 600-ohm input of the TL442 mixer IC.**

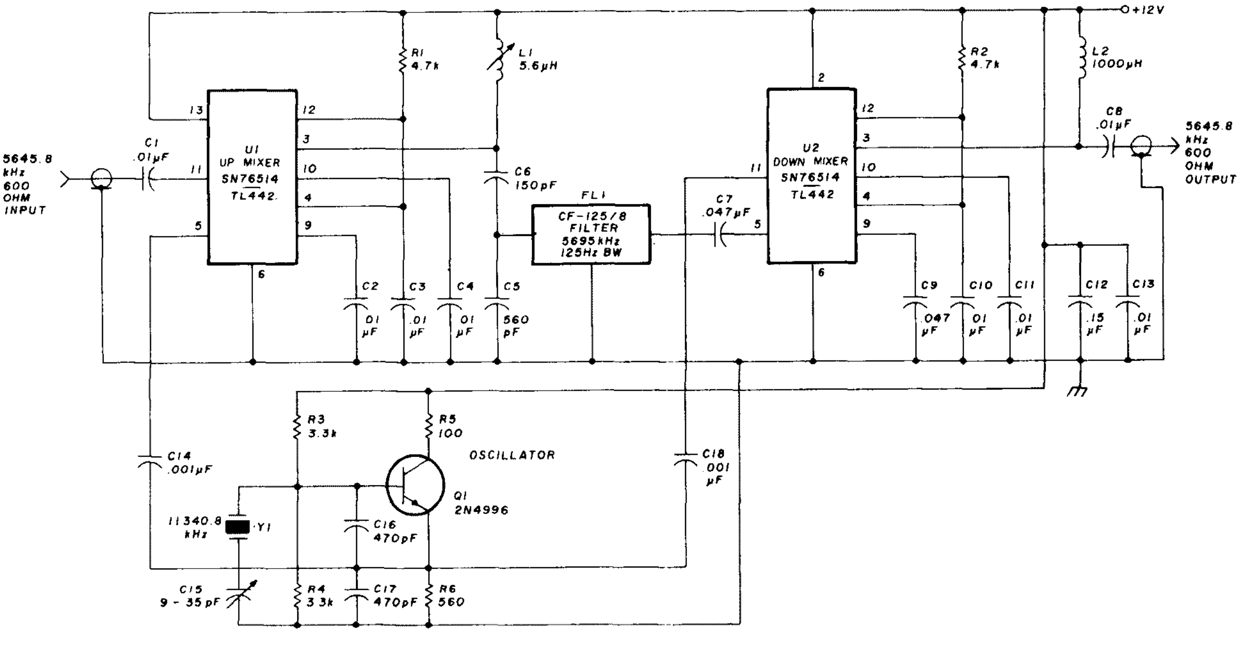



**fig. 5. Schematic diagram of the up/down converter. All resistors are ¼ watt, 10 per cent. The TL442 pin-out is shown for the dual in-line package. The active devices can be obtained from Texas Instruments Supply, 6000 Dentron Drive, Dallas, Texas 75235.**

*both* the input and the local oscillator signal. **Fig.** 3 shows a diagram of the Texas Instruments TL442 (old designation SN76514). This circuit was designed specifically for radio receiver applications. Its features include

**1.** Flat frequency response to 100 MHz; with tuning usable to 300 MHz, $C_i = 3\text{-}5$ pF; $C_o = 10$ pF

**2.** 50 and 600 ohms input impedances and 600/1200-output impedance

**3.** Factory-tuned null adjustments for both signal and local oscillator

**4.** Single- or double-ended voltage source

**5.** Differential amplifier with large signal-handling capability

**6.** Low-level local oscillator requirement

**7.** Noise figure of approximately 6 dB

**8.** Typical conversion gain of 14 dB

In the TL442 IC, uhf transistor chips are matched and the resistors are etch-trimmed in the manufacturing process to achieve balance. The IC actually consists of two cross-coupled differential amplifiers whose emitters are driven by a third differential amplifier. A constant-current source is connected to the third differential amplifier emitter. This device works best with 250 mV local-oscillator injection, and performs without significant overloading, up to about 300 mV of rf signal. Hence, the signal-handling characteristics of the TL442 are as good as or better than most vacuum-tube converters in current receiver designs.

An excellent description of the TL442 is also available from Texas Instruments.[1] Cost of the doubly balanced mixer is $2.40, an excellent trade-off when you consider that no external components are required. With more than 30 dB separation between the desired $f_{FL}$ signal and the nearby $f_{if}$ signal, the remainder of the high-performance crystal filter converter design is downhill.

Impedance matching, or the lack of it, is a big benefit of using the TL442. The fixed 600/1200-ohm output required no LC network, and only the most simple matching circuit to match the 50-ohm crystal filter. Going from the filter to the down-mixer does not require matching when using the 50-ohm input of the mixer! With isolation between the local oscillator and the output port of more than 30 dB, the local oscillator signal will have only minimal impact upon the receiver, and will provide more than 60-dB protection against signal leakage across the filter.

The gain/loss in the conversion process is also worth planning. The Sherwood CF-125/8 filter has a typical loss of 9 dB (maximum 11 dB). Another factor is the bandwidth reduction loss from 2.4 kHz to 125 Hz, which was shown to be about 13 dB. I like background noise to remain constant rather than to keep the signal strength constant when switching between the two filters; the noise floor is always a ready reference and a 13 dB drop in the noise floor is a noticeable deadening of the receiver! If the TL442 is connected for a 1200-ohm output impedance, about three S-units of excess gain can be provided to slightly more than account for loss of background noise due to bandwidth reduction.

Designing the matching network, from the TL442's 1200-ohm output impedance to the CF-125/8 crystal filter's 50-ohm input impedance, is based upon the capacitive tap-down network shown in **fig.** 4. The IC output impedance is 1200 ohms in parallel with 10 pF of source capacitance. L1 is used to resonate this 10 pF and the series connected tap-down capacitors, C1 and C2. **Fig. 4** shows the relationships between the network components and the termination parameters. As discussed before, one of the advantages of the TL442 is that the crystal filter will directly match the 50-ohm input of the second TL442 mixer.

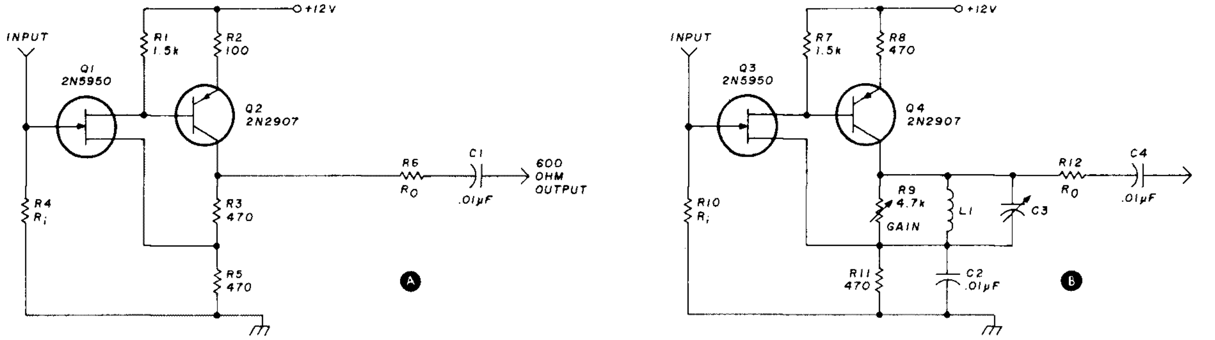



**fig. 6. The 1:1 line isolation amplifier is shown in *A*, while the amplifier with the variable ratio is shown in *B*. This design is capable of handling large signals with low cross-modulation.**

All that's needed to complete the converter are three semiconductor devices and one tuned circuit. **Fig. 5** shows the schematic. Both mixer ICs are configured in the same manner. Rf ground potentials are carefully bypassed with monolithic capacitors using short leads. The TL442 outputs are single ended, and the impedance is raised to about 1200 ohms by applying the supply voltage to pin 13. The constant-current source resistor network derives its voltage from the 3k resistor connected from pin 12 to pin 4; an additional 4.7k resistor is connected to pin 4 to increase gain. At the signal input, the 600-ohm input

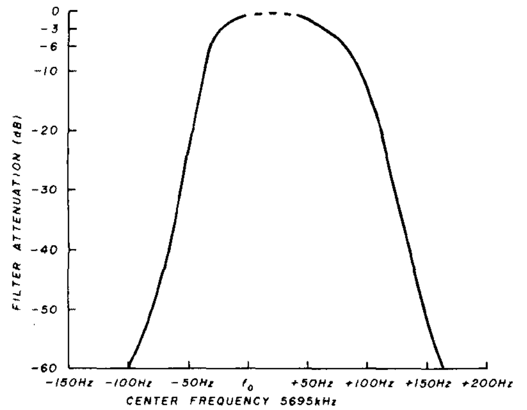



**fig. 7. Frequency response of the Sherwood Engineering CF-125/8 crystal filter.**

(pin 11) is used for the receiver first i-f signal because matching to 600 ohms is convenient.

The local oscillator was carefully designed to provide as much decoupling from the supply voltage as possible and also to provide a very low output impedance. A T-pad attenuator between the two mixers further decreases the possibility of signal leakage from the signal input through the input mixer, oscillator, and through the output mixer.

The capacitive tap-down network from the up-mixer, U1, to the crystal filter, FL1, is composed of L1, C6, and C5. Depending upon the *Q* of L1, and any other filter impedance, C5 and C6 can be adjusted to give the proper ratio for a good filter match. This occurs at or near maximum signal strength without objectionable ripple in the passband or out-of-band ripples.

Retune L1 each time C5 or C6 are changed; the adjustment is straightforward and noncritical. C15 provides a fine-tuning adjustment for the crystal.

## line amplifiers

When the output impedance of the receiver's first i-f is greater than one or two thousand ohms, a matching circuit will be required. Again, the capacitance tap-down network will work well for ratios of 24:1 or more. Further, there is sufficient gain in the TL442 IC mixers to recover a few dB of circuit loss. When excessive loss is encountered, a line amplifier (**fig. 6**) will help recover gain, or match extremely high-impedance circuits to low-impedance circuits. This circuit features several S-units of gain while exhibiting very large signal-handling capabilities with low cross-modulation distortion. The output impedance is 1200 ohms, untuned, and should be easy to match to the second-mixer circuit in any receiver.

## construction

The whole system was built on a double-sided printed circuit board that fits over the pins on the CF-125/8 crystal filter. The filter is securely grounded to the back plane of the printed-circuit board to reduce signal leakage around the filter. A piece of double-stick *Scotch* mounting tape was used to attach the entire up/down crystal filter converter assembly to an unused panel inside the receiver. One word of caution: always place a metal shield between the input wafer and the output wafer of the crystal filter switch to minimize signal leakage.

## results

True single-signal reception with the CF-125/8 crystal filter in tandem with an ssb filter is most gratifying. **Fig. 7** shows the frequency response of the CF-125/8 crystal filter by itself. Other crystal filters give equally impressive results. My R4-B receiver is equipped with a 1:1 line amplifier which drives the Drake ssb filters from the 2-crystal filter in the first i-f stage. The output from the ssb filter drives the line driver (adjusted to make up the 6 dB filter loss) and then the second mixer. The up/down crystal filter converter is switched in between the ssb filter and the second line amplifier. The tandem combination of filters does not ring, and 40 word-per-minute CW copy is possible. The noise and static effects that were so bothersome when a 125-Hz audio filter (shape factor 3) was used are now annoyances of the past.

If you are lucky enough to find crystal filters that are on the same frequency as your receiver's first i-f, all that is needed is a line isolation amplifier to buffer the filter, a matching network, and a line amplifier to make up the gain of the return signal.

## reference

1. *Balanced Mixer Application Note*, Section 6.6 SN76514/TL442, Linear Circuits Application Department, Mail Station 964, Dallas, Texas 75222.

**ham radio**

# how to select TTL sub-series ICs for different digital designs

The popular TTL family of digital ICs is widely used in amateur applications, but low-power, high-speed, and Schottky TTL have been largely neglected — here's how to select the best TTL sub-series for your own designs

**Through the years**, as the 7400 series of ICs has become the mainstay of TTL logic designers, more and more devices have been added to the family. In the last few years, both Fairchild Semiconductor and Texas Instruments have increased their commitment to the market by introducing expanded lines of high speed, low power, and Schottky-clamped devices. At the same time, extensive use of foreign production facilities has allowed a 50 per cent drop in prices, which distributors are now beginning to pass along to the consumer. Where does that leave you when you decide to build that new keyer or frequency counter? Consider the popular 7400 quadruple 2-input NAND gate, for example. There are five versions: the 7400, 74H00, 74L00, 74LS00, and 74S00. Which version is best suited for your purposes? What advantages does one version have over another?

Actually, each 7400 sub-series (H, L, LS, and S) has clear cut strengths and weaknesses which make the choice a lot easier than it may appear. The two major differences between each sub-series are speed (maximum operating frequency) and power consumption. In general, to gain speed, power consumption must be increased. This speed-power trade-off would probably settle the matter because you would pick the lowest power version that meets the required speed and stop right there, but a third factor comes into play: cost. To either increase speed or decrease power, the cost at least doubles over that of the standard (and the least expensive) version. A good rule of thumb is to use these special devices *only* when the standard 7400-series chips can't do the job.

## performance comparison

Let's go back to the 7400 quad 2-input NAND gate and look at the differences between each version and set down some general characteristics for each sub-series.

The 7400 typically operates from dc to 35 MHz, as will the remainder of the 7400 series. This is a *typical* specification and does not hold true in devices of higher complexity such as the Texas Instruments SN74144, which contains a BCD counter, a four-bit latch, and a BCD to seven-segment decoder-driver. The SN74144 is intended to be a one-chip replacement for the popular SN7490A counter, SN7475 latch, and SN7447 decoder-driver combination. Because of the high component density in the SN74144 (an equivalent of 86 gates on one chip), however, the typical counting frequency only extends to 18 MHz. Some of the earlier devices were equally slow. There was a time, not too long ago, when it was sometimes necessary to go through a handful of 7490 counters before you could find one that would work to 30 MHz. Therefore, it's wise to buy only devices with current date codes, unless the application isn't critical. Problems shouldn't pop up with any major suppliers, like those who advertise in *ham radio*, because the turnover is too high for 1973 chips to be still floating around in the open market.

Digressing a moment, the date code is a three- or four-digit number standardized by the EIA (Electronic

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Industries Association). It is stamped on every integrated circuit, usually, but not always, after inspection. Contrary to what you might think, you can get an untested IC with all the same markings as a first-rate unit. If the number has four digits, the first two represent the year of manufacture, such as 75, and the next two stand for the calendar week, such as 38, which would mean the thirty-eighth week of 1975 (the third week in September). If the number has three digits, the first is the year of manufacture (in the example above, the year would be cropped to 5), and the last two digits refer to the calendar week.

The typical low power 74L00 version will operate to 3 MHz, making it somewhat slower than the CMOS family. Every other sub-series is faster than the original type. Below is a list in terms of *typical* speed:

| | |
|---|---|
| 74L | 3 MHz |
| 74LS | 45 MHz |
| 74H | 50 MHz |
| 74S | 125 MHz |

These figures represent the highest typical clock rate for flip-flops. Once again, remember that each device must be considered on a one-by-one basis where speed is concerned, with the higher-density *units having lower maximum frequencies than their* less complicated brothers.

Power consumption is often compared using the power dissipation per gate for each series. This information is given in **table 1**, along with all other comparative figures, but in this case it is based on the average supply current, per gate, assuming a 50 per cent duty cycle. The average supply current data is readily available for individual devices, whereas power dissipation is generalized for all gates in the series. The difference in consumption will hold true, when comparing more complicated devices, so long as it is treated as an approximate ratio. In other words, if the average supply current for one gate of a 7400 is 2 mA, and the average supply current, per gate, of a 74L00 is 0.2 mA, then it is fair to say that *any* standard 7400 series device will require approximately ten times the amount of power than an L-series unit does. In practice, the actual ratio may be more or less.

To take several cases, the power ratio is 7.25 for the SN7490A, SN74L90 combination, and 13.15 for the SN7473, SN74L73 flip-flop pair. The average supply current values for the remainder of the devices are: 4.5 mA for the 74H00, 0.4 mA for the 74LS00, and 3.75 mA for the 74S00. Thus, if the 7400 is used to establish the standard unit of power consumption (2 mA = 1 unit), then the relative standings are 0.1, 0.2, 1.9, and 2.3 for the L, LS, S, and H sub-series, respectively.

## selecting a sub-series

There are three variables that must be considered when choosing the proper series: price, speed, and power. In general, the first and deciding requirement is that the chip will work up to the desired frequency. It is possible to approach the choice from a power consumption standpoint, but if power conservation is *critical*, it would be a good idea to see what can be done with a very low power series like the RCA CD4000 COS/MOS family. On a cost-effective basis the low-power 74L00 series is not as good as the COS/MOS family, which has a much lower power-cost product; COS/MOS will also work at higher frequencies (5 MHz for counters, 10 MHz for gates and flip-flops).

This brings us to the method for selecting the best sub-series once the speed requirements have been fulfilled: the *power-cost product*. Since both cost and power are to be minimized, it is easier to multiply the two figures together and deal with one variable instead of two. The lower the power-cost product, the more performance you get for your money. In practice, the product is calculated by multiplying the average current per gate (in mA) for the series, by the average increase in price of the series over that of the standard 7400 series (given as a multiple, such as 3.1 times cost). Data is provided in **table 1**. This is a method for standardizing the selection process, or a mathematical replacement for common sense.

As an example of how to use the chart, suppose you are planning to built a 10-MHz frequency standard. The 10-MHz specification puts everything in the running except the L series. If cost effectiveness is the object, a look at the lowest power-cost product reveals that the LS series is your best bet. Sheer

**table 1. Comparison of the various TTL sub-series showing clock, rate, power dissipation, propagation delay, relative cost, and power-cost product**

| series | maximum flip-flop clock rate | power dissipation per gate | gate propagation delay time | average cost increase over standard series | power-cost product |
|---|---|---|---|---|---|
| 74L00 | 3 MHz | 1 mW | 33 ns | 3.1 | 0.62 |
| 7400 | 35 MHz | 10 mW | 10 ns | 1.0 | 2.00 |
| 74LS00 | 45 MHz | 2 mW | 9.5 ns | 1.4 | 0.57 |
| 74H00 | 50 MHz | 22 mW | 6 ns | 1.9 | 8.49 |
| 74S00 | 125 MHz | 19 mW | 3 ns | 3.4 | 13.07 |

low cost, providing that a husky power supply is available, would be provided by a switch to the standard series. Not all device selections are that simple.

Let's assume you are designing a frequency counter. The goal is to build a model capable of counting to the highest frequency and requiring the least possible power, and using only the TTL series (no CMOS or ECL integrated circuits); price is no object. In case the design goal will not be met by using only one 7400 sub-series, the lowest power version having the necessary speed will be selected. Excess speed margins, when not needed, will be sacrificed for power conservation. Starting with the 10 MHz oscillator, choose a 7400. The 74LS00 and other sub-series chips have a reputation for not performing well in oscillator service. A key to this problem is the different biasing requirements for each sub-series. It is impossible to just simply remove a 7400 from an oscillator circuit and plug in a 74LS00 without changing external resistor values. There are many proven oscillator circuits based on the 7400, but little published information about biasing for oscillator service, so sticking to the well trodden path will assure success.

The first divider must be able to toggle up to 10 MHz; a 74LS90 will require the lowest power. All remaining dividers operate at 1 MHz or below, making the 74L90 the best bet. The counter control circuitry, which generates the count enable, strobe, and reset pulses functions at a very low rate, since most counters can make no more than 10,000 counts per second, so L-series devices can be used. The gate must pass the highest counted frequency, as must the first decade counter, and this application calls for Schottky ICs such as a 74S00 for the gate, and a 74S196 for the first counter. The second decade counter must be LS to count up to 12.5 MHz, but the remaining counters may be L versions. Latches and decoder drivers can also be chosen from the L series.

It is important to note that no standard series TTL logic was used in this circuit. Only when performance can be sacrificed in favor of price is standard TTL a wise choice. Price is almost always important, which explains my rule of thumb which suggests that the standard series be used exclusively, except when it just won't do the job. It's less expensive to build a higher current power supply than to purchase twenty-five special ICs at three times the cost of their standard TTL equivalents.

## TTL sub-series compatibility

One of the original design objectives for the different TTL sub-series was compatibility. All have the same maximum supply voltage rating of 7 volts, except for the L series, which is 8 volts. This gives plenty of leeway above the typical 5.0 V supply voltage, which is common to all sub-series. Operating temperature range extends from 0 to 70°C (32 to 158°F). The maximum input voltage for the L series is 7 volts, with 5.5 volts as the limit for all others. Because of these similarities, mixing devices from different sub-series will produce no problems so long as fan-out limits are observed.

Fan-out (the number of inputs a single output can drive) is figured only on the basis of outputs driving inputs from the same sub-series. The standard fan-out is 10 loads, except for the L and LS series, where it is 20. Mixing of devices is permitted as long as the output can source (provide) or sink (absorb) the *total* current to or from all inputs.

The high- and low-state input requirements are shown in **table 2** along with output sink capabilities.

**table 2. Input and output data for the various TTL sub-series. Note that the L series had two different standard inputs; assume highest input current when calculating output requirements. Negative signs represent current flow out of terminal**

| series | input current (high state) | input current (low state) | maximum output sink current | maximum output source current |
|---|---|---|---|---|
| 74L00 | 10/20 μA | −0.18/−0.8 mA | 3.6 mA | −200 μA |
| 7400 | 40 μA | −1.6 mA | 16 mA | −400 μA |
| 74LS00 | 20 μA | −0.4 mA | 8 mA | −400 μA |
| 74H00 | 50 μA | −2.0 mA | 20 mA | −500 μA |
| 74S00 | 50 μA | −2.0 mA | 20 mA | −1000 μA |

From these figures it's easy to check to see whether a particular output can handle its loads. Just add up the low-state currents for all loads (inputs), and then check that the totals fall within the maximum sink limit for the output. The negative values of input current represent a flow out of the terminal, back into the output. If the output can sink the required current, it will always be able to source enough current for the loads.

## conclusion

While the use of standard and LS series TTL ICs has certainly caught on for amateur projects, the H, L, and S series have been largely neglected. As so often happens with new products, this is due more to insufficient information than it is to a lack of applications. It is hoped that this article has provided enough information to generate more interest in using the various TTL sub-series ICs in future designs.

**ham radio**

# 500-watt regulated power supply

## Introducing Dragon One — a power source with built-in safety features for your high-power equipment

**With** solid-state devices **making** advances in the vhf and uhf spectrum, more powerful transmitters are made available to radio amateurs. Solid-state transmitters and linear amplifiers in the 150-200-watt range are not uncommon. Although most of the equipment is made to operate in automobiles using car batteries, many amateurs want to operate their higher-power solid-state devices in their homes. But here the power supplies become a problem, unless of course, you wish to purchase an auto battery and a battery charger as accessories for the modern equipment. Even then, the battery has a finite life, and the sulphuric acid in the battery is hazardous inside living quarters.

### the dragon one supply

This article describes a power supply system that delivers a whopping 500 watts of clean, regulated power. The regulation is better than 1 per cent from no load to full load, with a ripple voltage of less than 10 mV peak-to-peak. Safety features such as current limiting, overvoltage shutdown, and short-circuit protection are all built in. Optimum regulator efficiency is approximately 65 per cent at an output of 450 watts.

**Fig. 1** shows the regulated-power supply schematic. Transformer T1 steps down the line voltage to 22 volts, which is rectified by full-wave bridge rectifier CR1. Filtering is by C1, which is a computer-grade electrolytic capacitor having a capacitance of 18,000 $\mu$F. R1 discharges C1 after the power supply is turned off.

Regulation is provided by U1, the popular 723 regulator IC. The 12-15-volt voltage adjustment is by R4. C5 ensures oscillation-free operation of regulator

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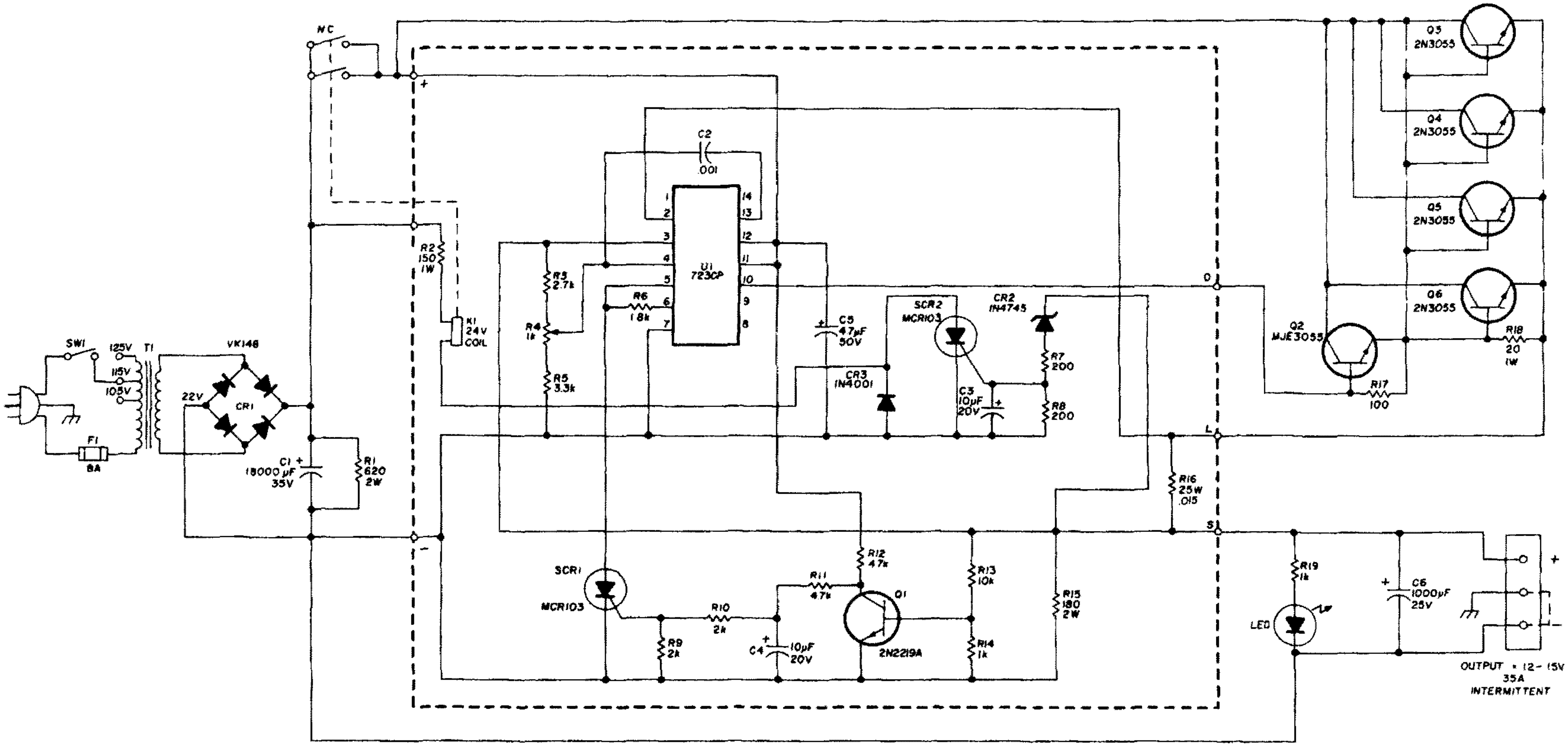



**fig. 1. Power-supply schematic. All resistors, unless otherwise marked, are ½ watt, 10 per cent.**

**table 1. Parts list for the 500-watt regulated supply.**

| | |
|---|---|
| CR2 | 16V zener 1W, 1N4745 or equivalent |
| CR3 | diode 50V 1A |
| LED | red light-emitting diode |
| T1 | transformer Lotronics T2230 |
| CR1 | bridge rectifier Varo VK148 or equivalent |
| C1 | electrolytic capacitor - 18,000 μF, 35V |
| Printed circuit board | Lotronics PC723SO |
| Heat sinks | 10″ x 6″ (254x152mm) Lotronics H10-6, one piece<br>4″ x 6″ (102x152mm) Lotronics H4-6, two pieces |
| R16 | resistor 15 milliohm Lotronics R15M |
| Chassis | 7″ x 8″ x 10″ (178x203x254mm) steel box |
| K1 | relay 24V coil dpdt, contacts rated at 10A 125V each |
| U1 | 723 regulator DIP package |
| SCR1, SCR2 | MCR103 or equivalent (50V 200μA gate current) |
| F1 | fuse holder and 8A fuse |
| SW1 | toggle switch dpst on-off |

Miscellaneous wire, screws, washers, terminal block, line cord.

The following parts are available from Lotronics, Box 975, El Cerrito, California 94530:

| | |
|---|---|
| Items 1-9 | Dragon One major component kit — $129.50<br>Dragon One complete component kit — $169.50<br>Dragon One assembled and tested — $209.50 |

For all above add $10 for shipping, insurance, and handling. California residents add 6 per cent tax. Instructions included.

723. U1 output drives Q2, an MJE3055, which in turn, drives Q3-Q6, 2N3055s. Current sensing is by R16, a special 15-milliohm resistor. The two output terminals are isolated from chassis ground. Grounding is achieved by connecting the positive or negative output terminal to the ground terminal with a jumper. A light-emitting diode indicates the presence of dc output voltage. R3, R4, and R5 make up the output voltage sensing divider; the voltage control signal is connected to U1 inverting input.

To protect the power supply from burning itself up in case of excessive load current, the short-circuit shutoff is done in conjunction with the current limiting provided by the regulator through R16. As load current exceeds 35 amps, the output voltage starts to drop. When the voltage drops below 8 volts, Q1 turns off and SCR1 turns on, pulling the regulator noninverting input close to ground potential, thus turning off the output power. This condition remains until the power supply is turned off and SCR1 unlatches.

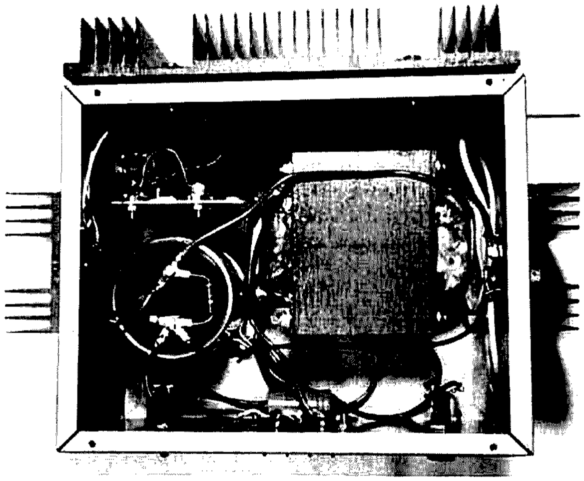
Overvoltage shutdown is designed into the system to protect your expensive transceivers and linear amplifiers. If anything should happen to the regulator or any of the series transistors, chances are one of these devices will short out, putting the full voltage across C1 at the output. This could be disastrous to transceivers and amplifiers. Relay K1, together with CR2 and SCR2 ensure that this will not happen, even if all the pass transistors and regulator are shorted. As the voltage exceeds 16 volts, CR2 starts to conduct, supplying gate current to SCR2, which turns on and activates K1. In doing so, the main dc supply is cut off and will remain off for as long as the power is on and the defect has not been corrected. This special feature is valuable and its additional cost is well justified, although the overvoltage shutdown feature may never be needed in the lifetime of the power supply.

## construction

All components are packaged in a 7 x 8 x 10-inch (178x203x254mm) steel chassis box. Three heatsinks are used (photo).

All components shown inside the dotted line in the schematic diagram are mounted on the printed circuit board. Since this circuit is a high-current power source, no. 12 (2.1mm) wire should be used for all high-current paths. However, no. 16 (1.3mm) wire can be used for interconnections from the two relay contacts, which are wired in parallel to the individual transistor collector and from the individual emitter to point L or R16. R19 and C6 are mounted behind the output terminal block. Holes are punched on the top and bottom panels for ventilation purpose.

Output voltage can be adjusted between 12-15 volts dc. Load current is rated at 35 amps intermittent, and 22 amps continuous duty. For prolonged operation at high current and low output voltage (below 13 volts), a small external fan is recommended for cooling the heat sinks. However, the power supply can deliver 22 amps continously without forced-air cooling if ambient temperature is below 77°F (25°C). The temperature of the pass bank tran-



**Underchassis view of the power supply. Three heatsinks are used. The heatsink mounted on the rear of the chassis box is isolated from chassis ground. The four 2N3055s (Q3 - Q6) are mounted directly on the heatsink. The heatsink on the right-hand side is for bridge rectifier CR1; the other heatsink is for Q2. Heatsink compound was used for mounting Q2 - Q6.**

sistor under this condition stabilizes at around 221°F (105°C). With a 25-30 cfm ($7 \times 10^5 - 8 \times 10^5$ $cm^3$/minute) fan blowing at the rectifier and the transistor heatsinks, the transistor heatsink temperature stabilizes at 122°F (50°C) with 30 amps continuous load current operation for one hour. Regulation is below 1 per cent from no load to full load (35 amps). The taps on transformer T1 are for optimum efficiency operation. It's obvious that if the line voltage is high, the unregulated dc voltage will also be high, making the voltage drop across the pass bank transistor high. That means higher power dissipation and lower system efficiency. Hence, if the input line voltage is connected to the proper tap, an optimum system efficiency is achieved.

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**ham radio**

# voice-operated gate

## to replace voice-operated relays for carbon microphones

Presenting a circuit using four ICs plus a couple of transistors and diodes to replace the old voice-operated relays in ssb transceivers

**VOR is an acronym** for what is often called the *"voice-operated-relay"* or "squawk-to-talk" circuit, as used in many modern ssb and fm transceivers. "Voice-operated-relay" was an adequate description when tubes and relay circuitry were used, but it's rather unusual to find such relays in today's all-solid-state designs. And so now we have the Voice Operated Gate, or VOG.

The VOG described here used four ICs plus a couple of transistors and diodes to accomplish pre-amplification, bandpass filtering, and audio gating. A logic output also comes out of the VOG (choice of **1** or **0** for *gate-on*), to serve as a turn-on signal for other sections of the system being voice controlled.

Incorporated in the VOG is a lowpass and highpass filter pair providing the equivalent of a 300-3000 Hz bandpass filter with 40 dB per decade rolloff at each edge. These filters are of the active type, built around operational amplifiers. Only the audio passing through the filters can actuate the gate (and thereby pass through the VOG); this helps to discriminate against ambient noise.

### VOG circuit

A diagram of the VOG is shown in **fig. 1**. The first section is a microphone preamp with an fet constant-current source for a carbon microphone. The carbon microphone is a variable resistance, so the injection of a constant-current into it causes the voltage across it to be representative of the variations in resistance of the microphone. The op amp that forms

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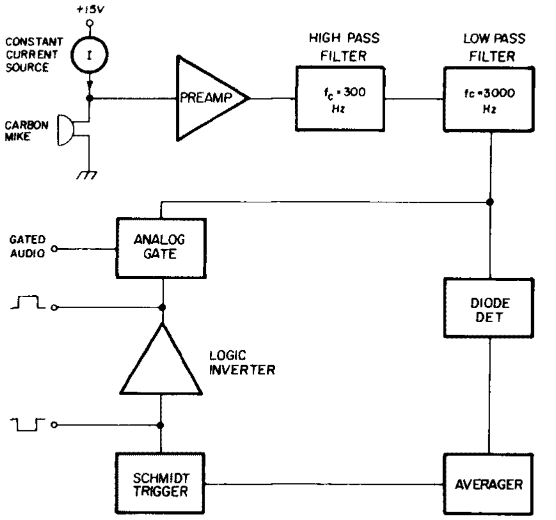


**fig. 1. Voice-operated-gate (VOG) circuit block diagram. The circuit is a replacement for the old voice-operated relay systems prevalent in many modern ssb and fm transceivers.**

the microphone preamp has a voltage gain of 100, which provides a voltage output of about 3 volts rms for usual carbon microphones during average close-talking use.

Following the microphone preamp is the highpass active filter followed by the lowpass active filter, each consisting of one section of the same quad op amp (U1) that's used as the preamp (see **fig. 2**). The last section of U1 is used as an *active* diode detector CR1, CR2, in which the op amp *linearizes* the detector. The diode detector is arranged to furnish the negative polarity of rectified audio.

The rectified audio from CR1, CR2 is then averaged by U2. Since the averager (U2) is also an inverter, the negative rectified audio is inverted and averaged to become a smoothed, long, positive pulse of the duration of the audio burst originally delivered by the microphone. This positive pulse is processed by U3, a Schmitt trigger, which sharpens the pulse leading and trailing edges and makes it

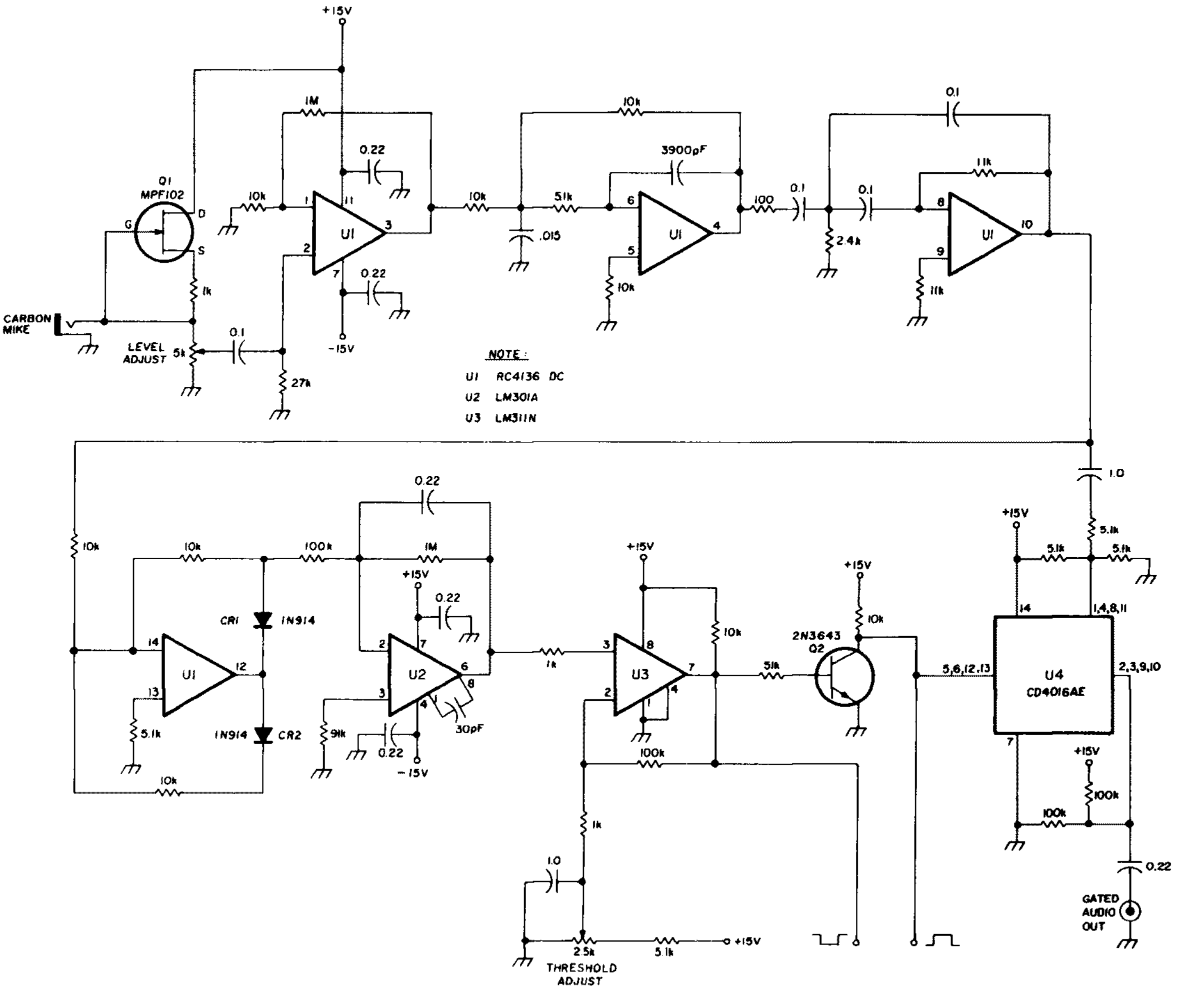


**fig. 2. Schematic showing the carbon microphone preamp, bandpass filter, and voice-operated gate (VOG). Note that U1 is the equivalent of four μA741 op amps and could be replaced by four such ICs.**

CMOS-logic compatible. The Schmitt trigger also inverts the pulse and adds an effect called hysteresis. That is, U3 output (pin 7) will go from **1** to **0** at an input level (set by the threshold-adjust pot) of say, 2 volts. U3 output will not return from **0** to **1** until the input voltage has dropped substantially *below* 2 volts. This hysteresis action prevents noise on the audio and minor voice level wavering from causing a chopping effect.

After the Schmitt trigger comes Q2, a simple transistor inverter, which inverts the audio-derived pulse to provide the proper polarity to turn on analog gate U4 when an audio signal is present. The inverter output and the Schmitt trigger output provide both **0** to **1** and **1** to **0** logic-level outputs, which can be used to actuate the turn-on function of the transmitter. Both polarities are handy, because this unit may be used with a number of transmitter designs.

The analog gate, U4, is a member of the RCA CD4000 CMOS logic family, which makes it much less expensive than some of the hybrid analog gates on the market. U4 consists of four analog gates. Since we need only one, all four sections have been wired in parallel. The CD4016 doesn't tolerate very large ac voltages without distortion, so the (filtered) audio input is attenuated at a ratio of 3:1 by a voltage divider at the analog input.

## adjustment and testing

Setup of the VOG is simple. Connect it to a carbon microphone and a ±15-volt supply. Connect a scope or ac VTVM to U1 pin 3 of U1. Talk into the microphone and adjust the LEVEL pot (**fig. 2**) until about 3 volts rms is seen, then adjust the THRESHOLD pot until about +2 volts is seen at its wiper arm. Connecting a scope or ac VTVM to the output should now show a pulse of audio when speaking into the microphone. A dc voltmeter at U3 pin 7 should jump from +15V for "no talking" to near zero for "talking". The same dc voltmeter at the Q2 collector should react in the opposite way: near zero for "no talking" and +15 volts for "talking."

## closing remarks

This VOG circuit was originally designed to replace one of the special-purpose ICs made by a large linear IC manufacturer. It surpasses the device it replaces in every way.

Note that U1 is the equivalent of four $\mu$A741 op amps and could be replaced by four such ICs. Also, two $\mu$A747s (dual $\mu$A741 op amps) or two MC1458s could also be used. U2 is best left as an LM301A, since the requirement here is for low input bias currents. When using other-than-called-for ICs, however, pin changes will have to be made.

**ham radio**

# accurate low power rf wattmeter for high frequency and vhf measurements

How to build an accurate low-power wattmeter that measures up to 10 mW from 1 to 500 MHz — it uses small lamps as barretters

**A pair of subminiature lamps** used as an rf power detector make up the heart of a simple but accurate rf power meter, which can be calibrated directly from dc measurements. The instrument described in this article can be used to accurately measure rf power from 10 mW down to about 0.2 $\mu$W, over a frequency range from 1 MHz to 500 MHz. Its high sensitivity makes it useful for a host of purposes including antenna gain measurements, local oscillator measurements, and vswr or filter response measurements in conjunction with low-power signal generators. Its maximum power capability can be extended to any level through the use of external calibrated attenuators or directional couplers. In addition, homebrew attenuators and directional couplers can themselves be calibrated using the power meter.

The rf power detecting element in the wattmeter consists of a pair of incandescent lamps used as barretters. Barretters have been used for many years in commercial wattmeters and have been discussed in several previous articles.[1,2]

A barretter is a wire element whose resistance increases with temperature. Suppose a barretter is heated to a specific resistance (say 50 ohms) by a variable power source whose level is known. As long as the total power dissipated and the ambient temperature remain constant, the barretter resistance will remain at 50 ohms. Now suppose the barretter is also heated with power from a separate source (an rf generator in this case) whose level is unknown. The resistance of the barretter will increase. If the power supplied from the known source is then reduced until the barretter resistance returns to 50 ohms, the amount of power reduction from the known source will equal the power supplied by the unknown source. The unknown power level is thus measured by metering the decrease in the known power source.

The known power source can be adjusted automatically to maintain constant barretter resistance by using a bridge circuit in a closed loop with an amplifier. In many commercial microwave power meters the closed loop forms a self-balancing audio oscillator so that the known power source is an ac signal (in combination with some dc which is also applied). The oscillator technique has the advantage of eliminating dc offset drift errors in the balancing and metering circuits. In the power meter described here, however, the known power source is pure dc. The dc approach was chosen for ease of calibration and testing, for circuit simplicity, and to allow a wide rf frequency range. (The relatively large rf coupling capacitor required for low-frequency response would introduce excess phase shift and upset the balance

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in an ac balanced bridge, depending on the low-frequency impedance of the rf source.)

## circuit description

The circuit diagram of the rf wattmeter is shown in **fig. 1**. The design philosophy was to explore what useful sensitivity could be achieved in a simple circuit without the use of special low-drift components or special schemes for drift compensation. The experimenter who wishes to build his own version of the wattmeter is encouraged to try his hand at improvements.

The incandescent lamps, I1 and I2, used in the rf sensor are subminiature T-3/4 types obtained at a hamfest flea market. The lamps have wire leads and the glass envelopes are 0.187 inch (5mm) long by 0.094 inch (2.5mm) in diameter. Their dc characteristics indicate they are similar to Chicago Miniature types CM2, CM30, or CM3102. **Fig. 2** shows the measured current-voltage (I-V) characteristic of one of the lamps. Note the non-linear nature of the plotted data which indicates changing lamp resistance. This general characteristic is typical of all incandescent lamps with tungsten filaments. **Fig. 3** shows the same data plotted as dc resistance, V/I, versus power dissipated, VI. At rf frequencies, the resistance of the lamp during any rf cycle remains constant and equal to the dc resistance because one rf cycle is much shorter than the minimum thermal response time of the lamp filament (skin effect does not appear to seriously alter the resistance of the small diameter, high resistivity filament over the frequency range of interest).


**The author's completed power meter.**

In order to simultaneously feed dc and rf to the lamps over a wide bandwidth, the lamps are connected in series for dc and in parallel for rf. Chip capacitors C1 and C2 perform the functions of rf coupling and bypassing, respectively, with low impedance over a wide frequency range. If chip capacitors are not available, small ceramic disks with zero lead length may be used. For good uhf measurement accuracy, construction of the rf sensor must be based on good uhf construction practices, with emphasis on minimizing parasitic inductances by keeping all leads short. The rf paths through C1, either lamp, and C2 to ground must be as short as possible.

The rf sensor is built on a small piece of double-clad glass-epoxy printed-circuit board 1/16 inch (1.5mm) thick as shown in the photograph. Both sides of the board were soldered directly to the rear of a BNC connector, with the connector center pin soldered to a pad approximately 0.105 inch (2.5mm) wide. This pad forms a 50-ohm microstrip transmission line leading to chip capacitor C1. One lead of both I1 and I2 is soldered to a small pad connected to the other end of C1. A small hole is drilled through the board at the ground lead of I2 so this lead can be soldered to the ground plane on both sides of the board. The opposite lead of I1 is soldered to the pad in the upper right hand corner in the photo. A dc feed wire is also soldered to this pad and chip capacitor C2 is soldered from the point of attachment of I1 across a gap to the ground plane.

On the ground side of C2, another hole is drilled through the board and a wire is soldered through this hole to form a direct connection to the ground plane on the back of the board. The use of small filament lamps with low parasitic inductance, and this method of construction ensure good performance into the uhf portion of the spectrum. (**Warning**: chip capacitor ends must be soldered quickly with minimum heat; otherwise tin-lead solder will rapidly leach away the metallization from the ends of the capacitors.)

The layout of the remaining dc portion of the wattmeter circuit is not particularly critical and was built on a Vector DIP padboard mounted on the meter terminals. The unit is housed in a 4 x 5 x 6 inch (10.2 x 12.7 x 15.2cm) minibox.

The lamps are operated at sufficient dc current to bring their series resistance to 200 ohms. If the lamps are reasonably well matched, the resistance of each lamp will be about 100 ohms, making the parallel rf resistance equal to 50 ohms. If two lamps identical to the one plotted in **fig. 3** are used, each will dissipate about 7 mW at a resistance of 100 ohms, for a total dissipated power of 14 mW. Thus, 14mW is the maximum rf power which can be measured in a 50-

ohm system with two such lamps. A highest scale of 10 mW was therefore chosen for the wattmeter. Random drift establishes a practical limit of 10 $\mu$W for the most sensitive scale.

To maintain their series resistance at 200 ohms, the lamps are operated in a bridge circuit consisting of R1, R2, R3, and the rf sensor. For best accuracy, R1, R2, and R3 should all be selected to be as close as possible to 200 ohms with R1 and R2 selected for best match, and R3 selected closest to 200 ohms.

The voltage difference between the two legs of the bridge is sensed and amplified by U1, a $\mu$A741 or similar type op-amp IC. The capacitors in the feedback loop of U1 form an integrator for very high dc gain and good stability. The output of U1 passes through diode CR1 to transistor Q1, the bridge current driver. Q1 is connected as an emitter follower, and supplies the necessary current to bring the bridge to a balanced condition. The 10k resistor across Q1 feeds a small residual positive bias to the bridge to ensure that the bridge will always come to balance with a positive potential, even though U1 may initially turn on with a negative output. Diode CR1 prevents emitter-base breakdown of Q1 if U1 turns on with a negative output.

Following turn-on, the output of U1 will quickly become positive in response to the residual positive bias on the bridge. The voltage at the output of U1 will continue to increase until enough current flows through the rf sensor to bring its resistance to 200 ohms, at which point equilibrium is achieved. In practice, the bridge comes to balance within a second or two of turn-on, with some overshoot due to the thermal lag of the lamps.

The equilibrium voltage at the top of the bridge, $V_B$, (3.50 volts in the unit shown) is fed to the metering circuit made up of U2A, U2B, and associated components. Range switch S2A selects one of the calibration resistors, R4 through R10. A method for calculating the values of these resistors is covered in the calibration section.

Op-amp U2A compares the voltage selected by S2A to a reference voltage established at pin 3 of its input. Since the full-scale voltage change in $V_B$ is only 1.1 mV for the 10 $\mu$W scale, the reference voltage supply must be extremely stable and minutely variable. To establish a stable reference voltage, fet Q2 is connected as a constant-current source feeding zener diode, CR2. Any fet having an $I_{DSS}$ of 3 mA or more could be used for Q2. Alternatively, a 5-volt, three-terminal regulator IC could probably be used instead of Q2 and CR2.

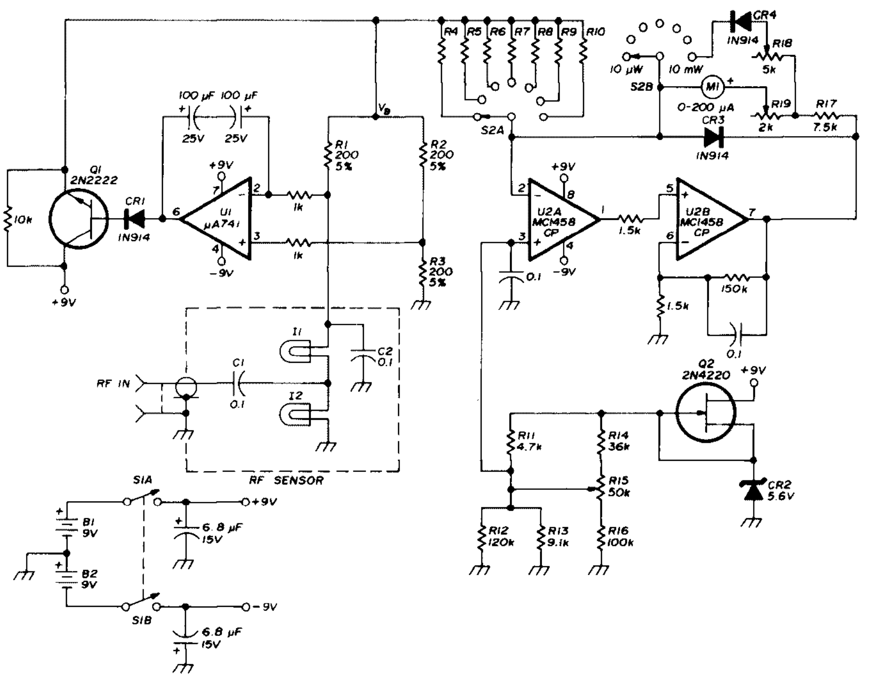
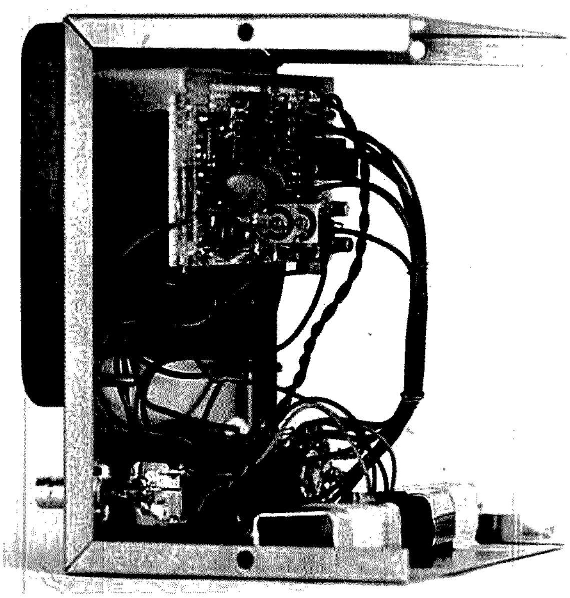



**fig. 1. Schematic diagram of the rf wattmeter for 1 to 500 MHz. Fixed-value capacitors are disk ceramic except as noted; polarized capacitors are electrolytic or tantalum; resistors are ¼ or ½ watt carbon composition types.**

| | |
|---|---|
| C1, C2 | 0.1 $\mu$F chip capacitor or miniature leadless ceramic discap |
| I1, I2 | subminiature T-3/4 incandescent lamp (Chicago Miniature type CM2, CM30, or CM3102) |
| J1 | BNC jack, flange mount |
| R15 | miniature 50k 10-turn pot |
| R18 | 5k trimmer |
| R19 | 2k trimmer |
| R4-R10 | (see **table 1** of text) |
| S1 | dpst toggle switch |
| S2 | 2-pole, 7-position rotary wafer switch |

Resistor network R11 through R16 divides the zener voltage down to the value required to match $V_B$. To get the required voltage resolution with a reasonable adjustment range, a miniature 10-turn pot was used at R15. If a 10-turn pot is not available, then both a coarse and a fine adjust pot must be used. Resistors R11 through R14 and R16 reduce the adjustment range of R15; this increases resolution. Resistors R11 through R13 are chosen to establish a reference voltage close to $V_B$ with the wiper of R15 disconnected. Resistors R11 through R13 also serve to maintain a fairly low impedance for the reference voltage. Resistors R14 and R16 are selected to reduce the adjustment range of R15, and to establish a residual voltage close to $V_B$ on the wiper of R15 when the wiper is set at mid-range.

Since the specified minimum open-loop gain of a single μA741 op amp is marginally low for proper operation of the metering circuit, two op amps are connected in cascade. Op amp U2B supplies an additional gain of 100 to the open-loop gain of U2A. A dual op amp, the MC1458CP, was used for U2A and U2B, though two μA741s could have been used or a quad 741 could have been used for the entire unit.



**Interior of the rf power wattmeter. All active circuits are installed on the perf board mounted on the meter terminals. The two incandescent lamps are mounted on the small section of printed-circuit board soldered to the BNC jack (lower left).**

The meter, M1, is connected in the feedback path of U2. Meter M1 is a 200 μA meter removed from an old vacuum-tube voltmeter. The action of U2 is to supply enough current through the feedback path to maintain the voltage at pin 2 of U2A equal to the reference voltage at pin 3. Since the current flowing in pin 2 of U2A is negligible, the current in the feedback circuit continues through the calibration resistor, $R_{CAL}$, selected by S2A. This current has no effect on $V_B$ because it is automatically compensated for by U1. By Ohm's law, the feedback current is equal to $\Delta V/R_{CAL}$, where $\Delta V$ is the difference between the reference voltage and $V_B$. On all scales except the 10 mW scale, all feedback current normally passes through meter M1. Diode CR3 conducts when the feedback current is negative, preventing M1 from pinning hard in the negative direction when the circuit is negatively unbalanced. Resistor R17 prevents M1 from being severely overloaded in the positive direction when the circuit is unbalanced positively. Resistor R17 is selected so that M1 reaches full scale somewhat before the output of U2B saturates in the positive direction. Resistor R18 and diode CR4 shunt some feedback current past M1 on the high end of the 10 mW scale to linearize the reading.

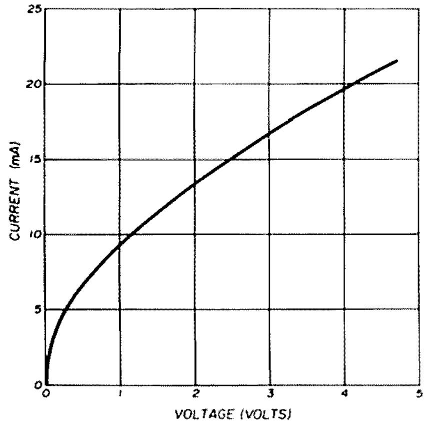



**fig. 2. Current-voltage (I-V) characteristic of an incandescent lamp of the type used in the rf power meter.**

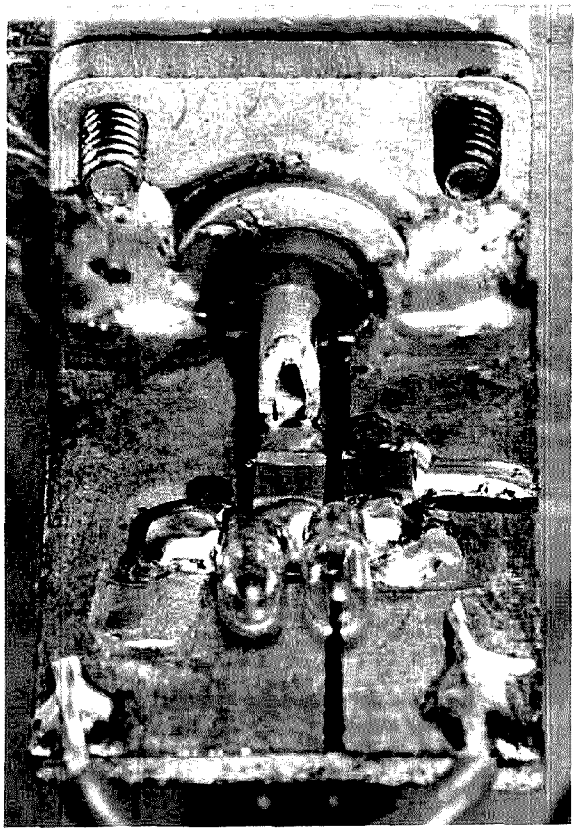
To allow portable operation, the unit is powered by two 9-volt batteries. Battery voltage *sag* following turn-on contributes some additional drift to the circuit. The miniature transistor radio batteries shown in the photograph sagged excessively and have been replaced by larger 9-volt batteries (Eveready 246). For enhanced stability, somewhat higher battery voltage could be used followed by electronic regulators to 9 or 12 volts. If it is desired to power the unit from the ac line, regulated dc supplies are a must.

The value of calibration resistance, $R_{CAL}$, for any scale is determined by calculating $\Delta V$, the change in $V_B$ for a given applied rf power level. The total dc power dissipated in the bridge is given by $V_B^2$ divided by 200 ohms, the series-parallel combination bridge resistance. Since each leg of the bridge has

equal resistance, the dc power dissipated in the rf sensor is 1/4 the total dc power dissipated in the bridge. The rf power applied to the sensor, $P_{rf}$, is equal to the difference in dc power dissipated in the sensor with no rf applied and the dc power dissipated in the sensor with rf applied, as expressed by

$$P_{rf} = \frac{1}{4}\,\frac{V_{BE}^2}{200} - \left(\frac{1}{4}\right)\frac{(V_{BE} - \Delta V)^2}{200} \qquad (1)$$

where $V_{BE}$ is the equilibrium voltage at the top of the bridge with no rf applied and $\Delta V$ is the change in bridge voltage following application of rf. Solving the above equation algebraically for $\Delta V$ results in the following solution:



Construction of the rf sensor showing the two incandescent lamps and chip capacitors C1, C2. Components are mounted on a small section of double-clad PC board which is soldered to the rear flange of the BNC connector.

$$\Delta V = V_{BE} - \sqrt{V_{BE}^2 - 800P_{rf}} \qquad (2)$$

A given desired full-scale rf power is used in **eq. 2** to determine a corresponding $\Delta V$. The required value of $R_{CAL}$ for proper full-scale reading is determined by dividing $\Delta V$ by $I_{FS}$, the full-scale value of meter current. **Table 1** shows the calculated values for the meter shown. Similar calculations should be made when duplicating the wattmeter, using the measured values of $V_{BE}$ and $I_{FS}$.

**table 1. Calculated values for calibration resistors for the rf power meter ($V_{BE}$ = 3.5 volts, $I_{FS}$ = 200 μA).**

| $P_{rf}$ | $\Delta V$ | $R_{CAL}$ |
|---|---|---|
| 10 μW | 1.143 mV | R4 = 5.715 ohms |
| 30 μW | 3.430 mV | R5 = 17.15 ohms |
| 100 μW | 11.450 mV | R6 = 57.25 ohms |
| 300 μW | 34.460 mV | R7 = 172.30 ohms |
| 1 mW | 116.200 mV | R8 = 581.10 ohms |
| 3 mW | 361.500 mV | R9 = 1808.00 ohms |
| 10 mW | 1.438 V | 7192 ohms (see text) |

The equation for $\Delta V$ is the equation of a parabola. Thus, the meter current varies parabolically instead of linearly with rf power. On the low-power scales, however, the voltage varies over such a small sector of the parabola that for all practical purposes it is linear. On the highest scale, the deviation from linear becomes significant, and is such that when the meter is calibrated for an accurate full-scale reading, the indicated power will be less than the actual applied power at levels below full scale. **Table 1** shows that a resistance value of 7192 ohms is needed for proper full-scale calibration of the meter on the 10 mW scale. For accurate calibration near the bottom of the 10 mW scale, a resistance 1000 times the value of the calibration resistor for the 10 μW scale, or 5715 ohms, would be required. Therefore, without some form of compensation, readings made near the bottom of the 10 mW scale will be only 79 per cent of the actual value, or 1 dB low.

To avoid lettering a special nonlinear 10 mW scale on the meter face, I used a compensation network. A compromise value of the calibration resistor R10 was selected at about 6000 ohms to reduce the error at the low end of the 10 mW scale. On the same scale, switch S2B connects the series combination of CR4 and R18 across M1 and R19. Toward the high end of the 10 mW scale, CR4 begins to conduct, shunting the excess current past M1. Variable resistor R18 determines the amount of current shunted away from M1, and variable resistor R19 determines the point at which diode CR4 begins conducting.

Before adjusting R18 and R19, an accurate voltmeter is connected from the top of the bridge (at $V_B$) to the reference voltage at pin 3 of U2A to read $\Delta V$. A value of $\Delta V$ corresponding to a full-scale reading of 10 mW (1.438 volt in the meter shown) is artificially established by adjusting the reference voltage level. Then R18 is adjusted for a full-scale

reading of the power meter. A $\Delta V$ corresponding to a reading of 6 mW (0.7705 volt in the meter shown) is then set and R19 is adjusted for a reading of 6 mW.

Since these two adjustments interact, they should be repeated several times until the meter reads both 10 mW and 6 mW. Linearization is now complete and the meter should be found to be quite accurate at all power levels.

The adjustment of R19 has no effect on the calibration of the other scales, provided the output of U2B is not at saturation for full-scale deflection of

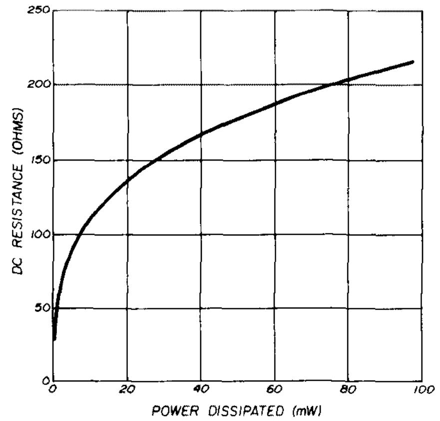



**fig. 3. Plot of lamp resistance vs power dissipated in the lamp.**

M1. Since the nonlinearity on the 3 mW scale is such that readings on the low end of this scale are only 5 per cent low (−0.23 dB), no linearization was deemed necessary for this and lower scales.

For most accurate results, the values of $R_{CAL}$ used for the lower scales should be as close to the calculated values as possible. Junk box resistors within a per cent or two of the desired values were selected using a digital ohmmeter. Where a proper value could not be found, a series or parallel combination was used.

The dB scale was added to the meter face so power could be read directly in dBm (dB with respect to a milliwatt) and so that losses and gains could be read out directly in dB. The scale position corresponding to each dB mark is given by

$$P = \frac{1}{antilog_{10}(0.1X)} \qquad (3)$$

where $P$ is the relative scale position (with 1 = full scale) and $X$ is the number of dB below full scale.

## procedure for use

Following turn-on, the meter is allowed to stabilize and the desired scale is selected. In the meter shown, stabilization is almost immediate on the higher power scales; several minutes are required on the 10 $\mu$W scale before warm-up drift ceases. Once the meter has sufficiently stabilized, the zero adjust pot, R15, is adjusted for zero reading. The rf power is then applied and readings are made. Provided the lamps are not burned out, the meter will not be damaged by exceeding the maximum power for the scale selected. Since the lamps can safely dissipate 200 mW, a considerable margin of safety exists. Random drift is significant on the 10$\mu$W scale; thus the meter zero should be checked between readings for greatest accuracy when using that scale.

## measured performance

Following calibration as described, the rf wattmeter was connected through one foot (30cm) of RG-58/U coaxial cable to the calibrated output of a Wavetek 3001 rf generator. Over the frequency range from 1 to 500 MHz, the generator power setting agreed to within 0.3 dB of the wattmeter reading at full scale on all wattmeter scales. The good agreement cannot be taken as a claim for wattmeter accuracy, however, because the specified worst-case generator power error on the most accurate power range is only 1.25 dB.

At 432 MHz, the input swr of the wattmeter was measured at 1.6:1. When measuring power from a 50-ohm source at 432 MHz, the resulting reading is calculated to be 0.24 dB low, due to reflected power. If the impedance of the source is adjusted to conjugately match the load presented by the wattmeter and interconnecting low-loss cable, this source of error is eliminated. On lower frequencies, the swr and resulting mismatch loss are expected to be even less because the parasitic reactance of the lamps and fixtures will be lower.

The wattmeter sees nearly constant use in testing rf circuits and devices of all types. Used directly, or with attenuators, it measures gains and losses. Used with directional couplers, hybrids, or rf bridges, it measures reflected power, return loss, and standing wave ratio. Since the lamps are a high temperature 50-ohm load, the wattmeter is also used as a noise generator for receiver rf amplifier tuneup and testing. In the few months since its construction, the wattmeter has become a virtually indispensable addition to my test bench.

## references


1. Robert S. Stein, W6NBI, "How to Use the Lab-Type Rf Power Meter," *ham radio*, April, 1977, page 44.
2. Bruce Clark, K6JYO, "RF Power Detecting Devices," *ham radio*, June, 1970, page 28.


**ham radio**

# drift-correction circuit

## for free-running oscillators

If you're bothered by warmup drift in your transceiver, here's a circuit that provides automatic compensation and uses readily available components

**The principle of drift** correction of an oscillator can be used in receivers or transmitters to compensate for warmup drift. The principle can also be used in new designs using simple free-running oscillators instead of the more complex types that use heterodyne mixing or phase-locked loops.

The idea is simple and straightforward. It can be best explained if you consider the operation of a frequency counter in which an oscillator frequency is measured. If the counter gate time is one second, and if sufficient displays are present, a 14-MHz signal could be displayed as 14.012.345 MHz. If, after the next measuring period, the least-significant digit changes from 5 to 7, for example, the oscillator frequency will have drifted 2 Hz high during that period. To counteract the drift, you could manually tune the oscillator back to its original frequency after each measurement. But there's a better way — read on.

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In the system described here, oscillator drift is compensated automatically. Only the last digit of the counter display is inspected after a measurement period. It is checked if the number is above or below a fixed value (5 in the example above). For values of 6, 7, 8, or 9, a voltage on a varicap in the oscillator reduces the frequency; for values of 0, 1, 2, 3, and 4, the reverse action occurs.

From this simple example it can be seen that:

**1.** The oscillator frequency always varies at a slow rate around a fixed value.

2. Stable points occur within 10 Hz from each other over the vfo tuning range.

3. Drift and short-term stability of the oscillator must be within limits. In the example cited, the drift must not exceed a few Hertz per second, otherwise the circuit can't compensate for the drift.

4. The automatic correction should be *very light*. If, after one correction period, the frequency overshoots too much, the remedy is worse than without the system.

For proper operation the correction-circuit time constant must be rather long (but also short enough to counteract the "natural" drift). Because of the long time constant, tuning feels quite normal. After a manual frequency adjustment, the frequency will creep to its nearest "stable" point (actually an unstable point) and will remain there. Because these points are closely spaced you don't notice the operation of the system by listening to a CW or ssb signal.

Note that, for correct operation of the system, the time base frequency doesn't have to be exactly 1 Hz, but the time base must be very stable. Thus the time base must be derived from a crystal oscillator. Counting can be in binary instead of binary-coded decimal format.

### circuit description

The circuit is shown in **fig. 1**. Only one stage of a counter is required. A 74LS93 binary counter (U1) counts the oscillator frequency that is to be stabilized. This stage is preceded by a 2N709 transistor

(Q1) to obtain sufficient sensitivity. About 100 mV of input signal is required.

After each counting period, the value of the $2^3$ output ($Q_C$, pin 8, of U1) is stored in a D-type flip-flop, U2, (half of a CD4013) at the rising edge of the time base signal. The flip-flop output drives an integrator (U3) up or down, which in turn drives a varicap in the oscillator to correct the frequency.

The time base frequency that actually determines system stability is derived by dividing the frequency of a crystal oscillator. A 1-MHz crystal oscillates with one input gate of a CD4060, (U4), which also contains 14 binary dividers. In combination with a CD4020, (U5), these two circuits divide the 1-MHz frequency by $2^{18}$ to about 3.81 Hz, so the stabilization points are spaced at 3.81 times 8 Hz, or 30.5 Hz.

I found that FT241 crystals between 400 and 500 kHz oscillate very well in this circuit. The total dividing factor should be $2^{17}$ in that case, which can be obtained by using output pin 2 of U4 instead of pin 3, as shown in **fig. 1**.

The counter counts almost continuously. Just after the transfer of the state of the $Q_C$ output to the D-type flip-flop (U2), a short reset pulse is generated by the other half of the flip-flop (U6). To achieve this action, the clock input signal of U6 is delayed by R1C1. After the Q output is set, the flip-flop resets itself because the Q output is connected through R2C2 to its own reset input. The resulting positive-going pulse is about 0.5 microsecond duration (line 3, **fig. 2**). This pulse resets the 74LS93 counter to zero which starts counting again immediately thereafter.

Worth mentioning is the long time constant of the integrator, which is formed by R3 and C3 (**fig. 1**). Capacitor C3 must be a low-leakage type, not an electrolytic. A polystyrene or polycarbonate type will do.

The switches labeled UP and DOWN (**fig. 1**) serve a dual purpose. First, after circuit switch-on, the integrator output can be brought into its range manually; but also, small frequency variations can be made by pushing the UP or DOWN button. So a push-button-controlled fine tuning is obtained, which is convenient if, for example, a CW signal slowly drifts out of a narrow CW-filter passband. (With this system installed you can be sure it's the other station that drifts.)

The CA3140, a very convenient operational amplifier, is used because of its high fet input impedance. The integrator output signal can be monitored on a meter to verify that it's still within its operating range. The action of the varicap in the oscillator must be such that a 10-volt output variation of the integrator shifts the frequency about 3 kHz.

## construction

The circuit was built onto a piece of *Vero* board and installed in my CW transceiver. A double-balanced diode mixer is used in my rig, so a high-level oscillator signal was available.

The UP and DOWN pushbuttons were mounted on the transceiver front panel. The control signal was monitored in a particular position of the transceiver meter switch.

Several prototype circuits were built using different construction methods, such as mounting all components on a copper-clad board with the ICs in sockets, but mounted upside down so that the socket pins could be wired directly. All these prototype circuits worked well, so the layout shown shouldn't be too critical. Just make sure that you avoid long wires between the ICs.

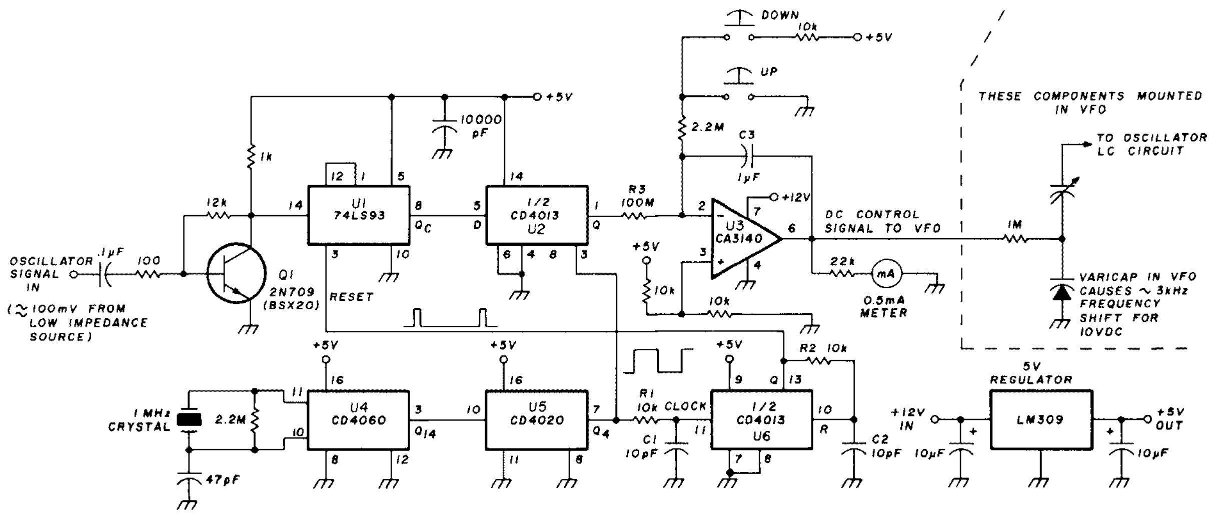


fig. 1. Circuit for vfo stabilization.

The circuit shown has been used for quite some time in my transceiver, which has a free-running oscillator on all bands. The highest frequency is 21 MHz, but the circuit has been used experimentally with oscillators operating to 40 MHz.

Within one minute after switch-on, the transceiver has crystal-quality stability on all bands. The 30-Hz frequency spacing between stabilization ponts is more than adequate for CW and ssb work. Also, during transmission, with about 200 watts to the antenna, a jump to another stabilization point has never occurred.

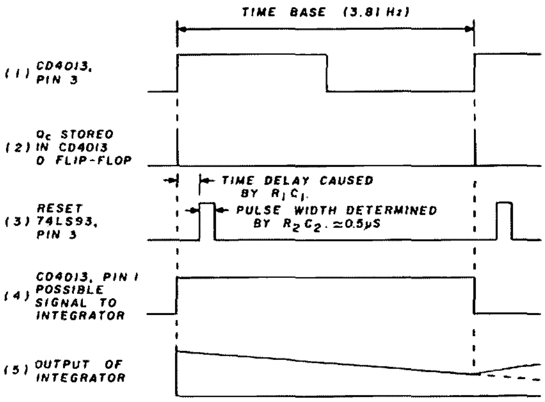


**fig. 2. Timing sequency of signals in the circuit of fig. 1. The time base is 3.8 Hz.**

A kind of proportional control system was tried instead of the constant-speed system described above. In this system, the corrective action depended on the *offset* value. Although the control could be measured (to be more effective), I believe this idea is really not worth the more complex electronic circuitry. Reason: with both systems a vfo becomes virtually drift-free, and both systems are not noticed during operation.

## conclusion

The system described here doesn't turn a bad vfo into a good one but helps to make a good one even better. Especially where a low-noise oscillator is important, as for local oscillators in high dynamic-range front ends for receivers to obtain low reciprocal mixing, I believe this technique could be applied successfully at least for hf-band applications.

Synthesized oscillators appear to be noisier than good free-runnng types so if this system is used in combination with a digital frequency read out, on a well-designed, free-running oscillator, a much simpler system results than is possible with fully synthesized oscillators, giving at least the same or better results.

**ham radio**

# active bandpass filters — some staggering thoughts

## Here's a rundown on stagger-tuned filters using op amps as active devices — great idea for many amateur applications

**Applications for active bandpass filters** in the audio-frequency range can be found in every part of amateur radio. Audio selectivity for CW, speech processing for ssb, tone-detector filters for RTTY, and control-tone separation for fm repeaters are only a few of the uses. In this article you'll learn an easy way to design and build stagger-tuned operational-amplifier active filters to fit your requirements. All you need is one of the readily available hand-held scientific calculators (or some other method for calculating square roots and logarithms).

Perhaps you've seen other types of active *filters* or filter designs using LC components. Why use stagger-tuned filters, and why use active filters? It's easy to build very narrowband audio filters by cascading, one after another, several identical simple filter sections. This may be adequate for some tasks but can often leave a lot to be desired in terms of transient response (*ringing*), peaked or *narrow-nosed* amplitude response, and poor skirt selectivity (*shape factor*). Conventional circuits using inductors can give excellent performance if well designed, which is often done with complex computer-aided design programs. But inductors are often large and hard to tune. Many amateurs have been discouraged by the need to add or remove turns from the 88-mH toroidal inductors common in RTTY use.

### features

The filters described here offer many advantages. They give amplitude response with flat or slightly rippled characteristics in-band. Out of band, they have excellent skirt selectivity and a shape factor that improves directly as more filter sections are added. As a bonus, the transient response is usually much better than narrow-nosed filters. Best of all, each stage can be tuned separately with no measurable interaction or detuning of the other stages — this is a real plus for experimenters.

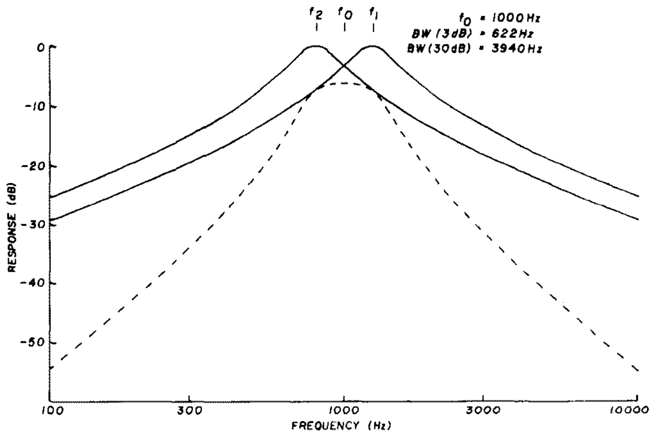


**fig. 1. Typical stagger-tuned response. By choosing the correct peak frequency, $f_n$, and $Q$ for each stage we get the response shown.**

Now the bad news (which isn't really too hard to take). Stagger tuning requires that each stage provide enough gain so that the sum of the stage gains is *greater than that of the overall filter*. This is because of *staggering loss*, of which you'll see more shortly. With op amp ICs and their large open-loop (no feedback applied) gain, this parameter turns out to be of little concern. Another problem is that if one

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of the stages is out of tune, the filter response can be poorer than in designs that purposely introduce interaction between filter sections, as in most LC designs or in *leapfrog* active filters, which are much more difficult to design.

## description

The stagger-tuned filter is made of two or more stages, each having a different peak frequency, $f_n$, with an associated $Q$ (which *may* be the same as the $Q$ of *one* of the other stages), and a certain amount of gain, $G$. The sum, in dB, of the gains versus frequency can be arranged to give a flat response over the band of interest. **Fig. 1** shows how this happens. In the area between the two peaks one response rises as the other falls. By choosing the right $f_n$ and $Q$ for each stage, we get the response shown. Note that, at the center frequency of the overall response $f_O$ (where the stages have equal loss), the net loss is *twice* as much (in dB). This is the stagger loss, $S$, which must be made up by the sum of the individual stage gains to give unity gain overall. Compare **figs. 1** and **2**. **Fig. 2** shows a two-stage nonstaggered or "synchronously tuned" filter response with the same 3-dB bandwidth. Note the poorer skirts and the much rounder passband.

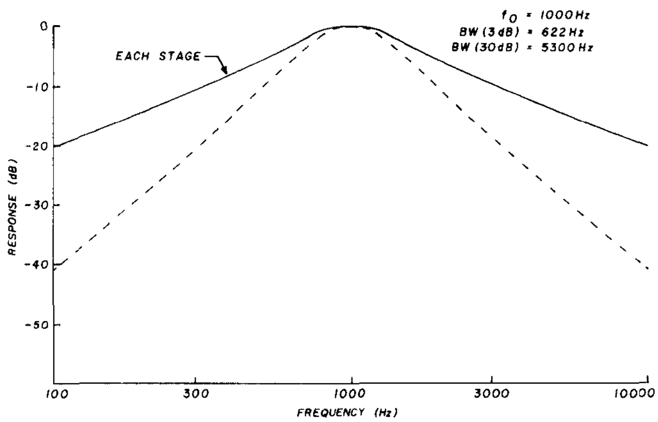


**fig. 2. Response of a two-stage synchronously tuned filter (compare with the response in fig. 1).**

In the stagger-tuned filter, the shape of each stage response is of the classic single-resonator shape (the same as that generated by a single parallel LC circuit with a shunt resistance to define the $Q$). The amplitude response is defined mathematically (for those of you itching to use your HP-25) as follows:

$$\frac{V_{out}}{V_{in}} = -\ 10 \log_{10} \left[1 + Q^2\left(\frac{f}{f_n} - \frac{f_n}{f}\right)\right]^2 \qquad \textbf{(1)}$$

**Eq. 1** is of interest only and is not necessary for designing a filter. It can be used for analysis, however. You can find the response of each stage then add all responses together to find the overall filter response. One thing that's important to note is that the curve has *geometric symmetry*. All this means is that if the upper (x) dB-down point is two times the center frequency, then the lower (x) dB point will be at one-half $f_n$. This relationship is expressed by

$$f_n = \sqrt{f_L f_H} \qquad \textbf{(2)}$$

where $f_L$ is a frequency below $f_n$ with the same attenuation as $f_H$, which is higher than $f_n$. Note that $f_n$ is *not* the arithmetic average of $f_L$ and $f_H$. The resultant overall filter response will exhibit the same type of symmetry as the stages of which it is composed. So **eq. 2** holds for the complete filter, where $n$ is zero.

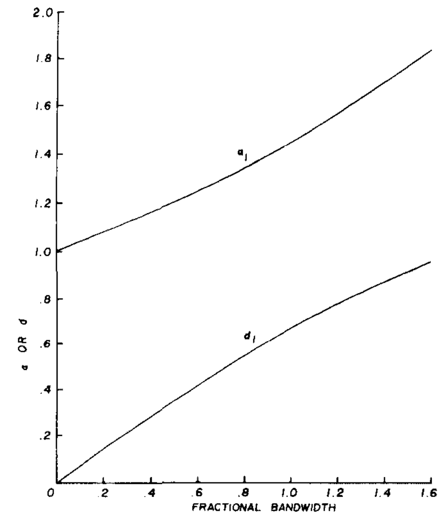


**fig. 3. A two-stage Butterworth filter showing $\alpha$ or $d$ as functions of fractional bandwidth, $\delta$.**

**Design procedure.** In designing the filter, the first thing is to decide what type of filter is wanted. The Butterworth, or maximally flat filter, provides the flattest passband and a good skirt shape. The Chebychev or equal-ripple filter gives ripples in the passband (1 dB in the designs to follow), but in turn, it has very rapid cutoff of the band. Many other filter types are in use, but these two will serve you well.

Next you must determine how many stages you want. This requirement is determined by the required shape factor, with the other consideration being how much circuitry you want to build. High $Q$s and more precise tuning of the stages are also requirements of the higher-performance designs.

**Shape factor.** To refresh your memory, shape

**table 1. Shape factors for Butterworth and Chebychev filter designs.**

| number of stages | butterworth 3/30 dB | butterworth 6/60 dB | 1-dB chebychev 3/30 dB | 1-dB chebychev 6/60 dB |
|---|---|---|---|---|
| 1 | 31.60 | 577.00 | 31.60 | 577.0 |
| 2 | 5.62 | 24.00 | 4.70 | 20.8 |
| 3 | 3.16 | 8.33 | 2.30 | 6.6 |
| 4 | 2.37 | 4.90 | 1.75 | 3.5 |
| 5 | 1.99 | 3.57 | 1.60 | 2.5 |
| 6 | 1.78 | 2.89 | 1.30 | 2.0 |

factor (also called *selectivity ratio*) is the ratio of bandwidth at a higher attenuation to the bandwidth at a lower attenuation. Most common is the 6 - 60-dB

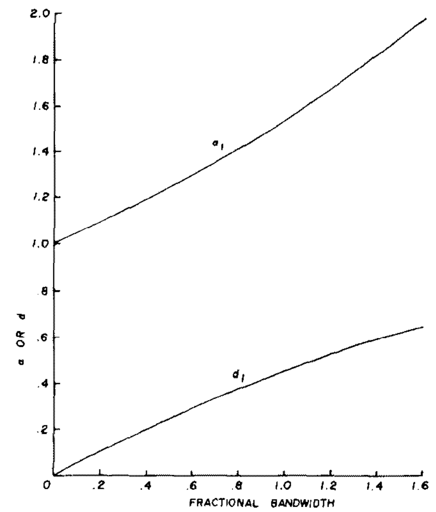


**fig. 4. Three-stage Butterworth filter showing $\alpha$ or $d$ versus fractional bandwidth, $\delta$.**

shape factor. **Table 1** shows this shape factor versus the number of stages for Butterworth and 1-dB-ripple Chebychev filters. Also given is the 3 - 30-dB shape factor.

After deciding the type and complexity of the filter, specify the lower 3-dB point, $f_L(3\ dB)$, and the

**table 2. Approximations for fractional bandwidth, $\delta$, equal to or less than 0.3 for Butterworth and Chebychev active filters.**

**Butterworth**

two-stage

$\alpha_1 = 1 + 0.365\delta$ $\qquad$ $d_1 = 0.707\delta$

three-stage

$\alpha_1 = 1 + 0.450\delta$ $\qquad$ $d_1 = 0.500\delta$

four-stage

$\alpha_1 = 1 + 0.485\delta$ $\qquad$ $d_1 = 0.380\delta$

$\alpha_3 = 1 + 0.195\delta$ $\qquad$ $d_3 = 0.920\delta$

**Chebychev (1-dB ripple)**

two-stage

$\alpha_1 = 1 + 0.365\delta$ $\qquad$ $d_1 = 0.433\delta$

three-stage

$\alpha_1 = 1 + 0.450\delta$ $\qquad$ $d_1 = 0.220\delta$

upper 3-dB point, $f_H(3\ dB)$. Then use **eq. 2** to find $f_O$. Next find $\delta$, the fractional bandwidth.

$$\delta = \frac{(f_H - f_L)}{f_O} \qquad (3)$$

This parameter, $\delta$, is the main design factor. It's used to find tuning data for each stage. Refer to **figs. 3, 4, or 5** for Butterworth filters of two, three, or four stages respectively. For a 1-dB Chebychev filter of

**table 3. Design equations for two-, three-, and four-stage filters. Parameter $\alpha$ is the ratio of resonant to filter center frequency.**

for two-stage filters

$f_1 = (f_0)(\alpha_1)$ $\qquad$ $f_2 = f_0/\alpha_1$

$Q_1 = 1/d_1$ $\qquad$ $Q_2 = 1/d_1$

$G_1 = G_2 = (S + G_0)/2$

for three-stage filters

$f_1 = (f_0)(\alpha_1)$ $\qquad$ $f_2 = f_0$ $\qquad$ $f_3 = f_0/\alpha_1$

$Q_1 = 1/d_1$ $\qquad$ $Q_2 = 1/\delta$ $\qquad$ $Q_3 = 1/d_1$

$G_1 = G_2 = G_3 = (S + G_0)/3$

for four-stage filters

$f_1 = (f_0)(\alpha_1)$ $\qquad$ $f_2 = (f_0)(\alpha_3)$ $\qquad$ $f_3 = f_0/\alpha_3$ $\qquad$ $f_4 = f_0/\alpha_1$

$Q_1 = 1/d_1$ $\qquad$ $Q_2 = 1/d_3$ $\qquad$ $Q_3 = 1/d_3$ $\qquad$ $Q_4 = 1/d_1$

$G_1 = G_2 = G_3 = G_4 = (S + G_0)/4$

two or three stages, see **fig. 6** or **7** respectively. From the appropriate figure, obtain $\alpha_1$ and $d_1$ (and $\alpha_3$ and $d_3$ for a four stage Butterworth). If your filter has a $\delta$ $0.3$, **table 2** offers approximations for $\alpha$ and $d$, which usually give better accuracy than reading from the graph. Decide what overall gain, $G_O$, in dB you want from the filter, then use **table 3** to find the tuning frequency, the $Q$, and the gain for each stage.

It's a good idea to organize the stages as given, with the highest-frequency stage first. (This mini-

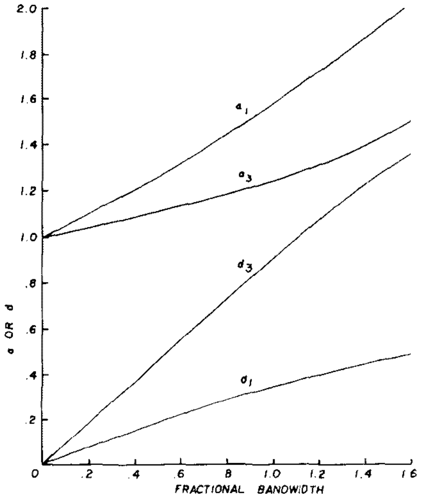


**fig. 5. Four-stage Butterworth, with $\alpha$ or $d$ as functions of fractional bandwidth, $\delta$.**

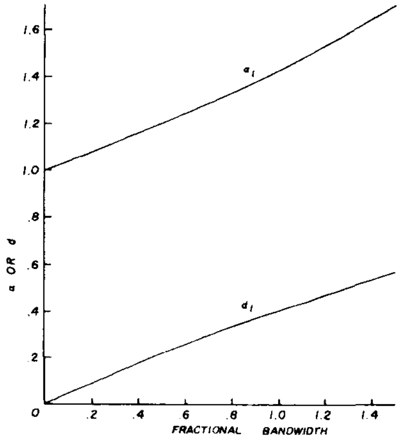


**fig. 6. A Chsbychev two-stage filter (1-dB ripple) showing $\alpha$ or $d$ as functions of fractional bandwidth, $\delta$.**

mizes harmonic distortion for the overall filter.) The higher-frequency stages have the lowest open-loop gain, which means that feedback will be less effective in reducing the distortion in these circuits than in the lower-frequency stages. Putting the low-frequency stages last gives maximum attenuation to any harmonics generated by the higher frequency stages.

**Multiple-feedback circuit.** Now that you know what the stages must do, the only thing remaining is to design circuits with the required $f_n$, $Q$, and $G$. For stages with low $Q$ (less than 10), the multiple feedback (MFB) circuit in **fig. 9** performs well. Almost any op amp will work here, but depending on its bandwidth, limitations exist on maximum $Q$ and maximum $f_n$.

The upper limit on $Q$ for the MFB circuit is given by the smaller of

$$Q_{max} \cong \sqrt{f_T/(5f_n)}, \qquad \textbf{(4)}$$
$$Q_{max} \cong 10$$

where $f_T$ is the frequency at which the op-amp gain equals zero dB (unity gain). The frequency, $f_n$, should be limited to about 1 per cent of $f_T$ (10 kHz for a 1-MHz $f_T$ amplifier, such as the type 741).

These restrictions minimize the effects of amplifier gain on $f_n$ and $Q$, which ensures accurate calculation of these parameters and freedom from drift because of amplifier gain changes with temperature.

The component values in the MFB circuit can be found easily. Choose convenient value of capacitor, $C$. The resistors are:

$$R3 = \frac{Q}{\pi f_n C} \qquad \textbf{(5)}$$

$$R1 = \frac{R3}{2 \cdot 10^{(G/20)}} \qquad \textbf{(6)}$$

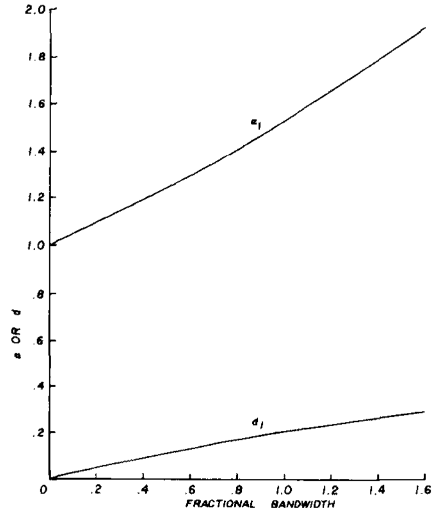


**fig. 7. Chebychev filter with three stages (1-dB ripple) showing $\alpha$ or $d$ versus fractional bandwidth, $\delta$.**

$$R2 = \frac{1}{[(2\pi f_n C)^2 R3 - (1/R)]} \qquad \textbf{(7)}$$

Note that $[10^{(G/20)}$ equals $antilog_{10}(G/20)]$, where $G$ is the gain, as described previously.

**State-variable design.** The limitations of the MFB circuit require that a higher-performance circuit be used in some cases. The state-variable circuit in **fig. 10** can do some amazing things. It can provide very high $Q$s (over 100) and is hard to beat for stability and lack of sensitivity to passive component drift. However, it does take two more op amps and four more resistors than the MFB design.

There are several degrees of freedom in this design. Choose $C$, $R2$, and $R4$ for convenience.* The remaining resistors are found from

$$R1 = R_2 Q/(10^{G/20}) \qquad \textbf{(8)}$$

$$R3 = \frac{1 - \dfrac{f_n}{f_T}}{2\pi f_n C} \qquad \textbf{(9)}$$

$$R5 = \frac{R4}{\left[\dfrac{2Q + (10^{G/20})}{\left[1 + \dfrac{4Q + (10^{G/20})}{\dfrac{f_T}{f_n}}\right]}\right] - 1} \qquad \textbf{(10)}$$

*A "convenient" capacitor is one as small as possible that doesn't require overly large resistors. Choosing resistors too much above 100k (for 741s or similar op amps) can lead to excessive dc offsets because of input-bias currents. Fet input op amps have extremely small bias currents and will tolerate resistors in the tens of megohms.

For the MFB circuit, capacitors with about 10 kilohms of reactance are in the ballpark. For instance, at 1500 Hz, a 0.01 $\mu$F capacitor is suitable. In the state-variable circuit, capacitors of about 100k ohms of reactance can be used, such as 0.001 $\mu$F at 1500 Hz. A reasonable value for $R2$ or $R4$ is between 10k - 100k.

Both circuits can be impedance-scaled if the calculations of component values reveal one or more values that are out of the desirable range. This means that all resistor values may be changed so long as all change by the *same ratio*, and the capacitors change by the *reciprocal* of that ratio. For example, if you find a 300k resistor where you'd like to have 100k, you can change it by making all the resistors one-third of their original value and by making the capacitors three times as large. In the state-variable circuit, $R4$ and $R5$ may be changed independently of the other resistors so long as the ratio R4:R5 is constant.

## design example

The design procedure is used to create an input prelimiter filter for an RTTY demodulator (TU). We'll

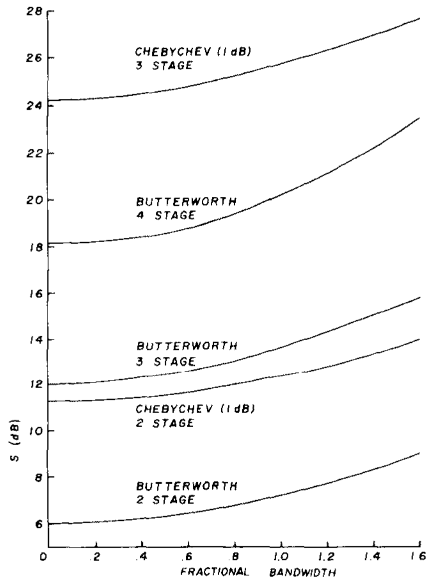


**fig. 8. Loss due to staggering, $S$, as functions of fractional bandwidth, $\delta$, for various active filters.**

choose an overall gain of 30 dB to provide adequate drive to the limiter from normal speaker signal levels. To give flat response in-band and reasonable delay distortion (associated with the transient response), we'll choose a Butterworth design. For good selectivity a four-stage configuration will be used. For 170-Hz shift and 45.45 Baud (standard 60 wpm), the

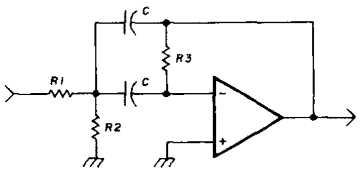


**fig. 9. Schematic showing the MFB, or multiple-feedback circuit.**

*CCIR* formula shows the bandwidth to be 246 Hz. To allow for tuning error and drift, a 300-Hz bandwidth at the 3-dB points will be used. The mark frequency is 2125 Hz; the space frequency is 2295 Hz. Thus the passband should be from $f_L = 2060$ Hz to $f_H = 2360$ Hz. **Eq. 2** gives

$$f_o = \sqrt{(2060)(2360)} = 2205\ Hz$$

From **eq. 3** we obtain the fractional bandwidth

$$\delta = \frac{(2360 - 2060)}{2205} = 0.1361$$

Since $\delta$ is less than 0.3, use the approximations in **table 2.**

$$\alpha_1 = 1 + (0.485)(0.1361) = 1.066$$
$$d_1 = (0.38)(0.1361) \quad = 0.0517$$
$$\alpha_3 = 1 + (0.195)(0.1361) = 1.0265$$
$$d_3 = (0.92)(0.1361) \quad = 0.1252$$

From **fig. 8** the loss due to staggering, $S_{,} = 18.2$ dB and from **table 3** we have

$$f_1 = (2205)(1.066) = 2351\ Hz \qquad Q_1 = 1/0.0517 = 19.3$$
$$f_2 = (2205)(1.0265) = 2263\ Hz \qquad Q_2 = 1/0.1252 = 8.0$$
$$f_3 = 2200/1.0265 = 2148\ Hz \qquad Q_3 = 1/0.1252 = 8.0$$
$$f_4 = 2205/1.066 = 2068\ Hz \qquad Q_4 = 1/0.0517 = 19.3$$

and $G = (18.2 + 30)/4 = 12.05\ dB$ (per stage).

It's apparent that the state-variable circuit must be used for the first and fourth stages ($Q >$ *than 10*). At 2351 Hz a 741-type op amp is capable of

$$Q_{max} \approx \sqrt{\frac{10^6}{5(2351)}} = 9.2$$

Since the second and third stages have $Q$s less than this, the MFB circuit is usable.

Let the capacitors in the state-variable stages be 0.001 $\mu$F and the capacitors in the MFB stages be 0.01 $\mu$F. Let R2 and R4 be 100k in stages 1 and 4. From **eqs. 8, 9,** and **10** we obtain the values for the first state-variable stage:

$$R1 = (100k)(19.3/10^{12.05/20}) = 482k$$

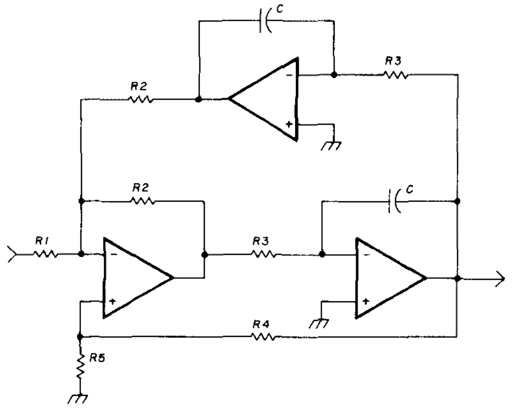


**fig. 10. Schematic of the state-variable stage.**

$$R3 = \frac{1 - \left[\frac{2351}{10^6}\right]}{(2\pi)(2351)(10^{-9})} = 67.54k$$

$$R5 = \frac{10^5}{\left[\frac{(2)(19.3 + 10^{12.05/20})}{1 + \frac{(4)(19.3 + 10^{12.05/20})}{\frac{10^6}{2351}}}\right] - 1} = 2686 \ ohms$$

Similarly, for stage four, we find

$$R1 = 482k \ ohms$$
$$R3 = 76.8k \ ohms$$
$$R5 = 2626 \ ohms$$

Now for stage two, using eqs. **5**, **6**, and **7**,

$$R3 = 8/\pi(2263)(0.01 \times 10^{-6}) = 112.5k \ ohms$$

$$R1 = 112.5k/(2)(10^{12.05/20}) = 14.05k \ ohms$$

$$R2 = \frac{1}{[2\pi(2263)(0.01 \times 10^{-6})]^2 \left[112.5k - \left(\frac{1}{14.05k}\right)\right]} = 439.6 \ ohms$$

And in the same fashion, for stage three, we have

$$R3 = 118.6k \ ohms$$
$$R1 = 14.8k \ ohms$$
$$R2 = 462.9k \ ohms$$

## construction

The filter was constructed using two MC3303 quad op amps. Combinations of one per cent resistors were used to give the calculated values within 0.5 per cent or less, nominally. Polystyrene capacitors, 1 per cent tolerance, were used in all sections.

The measured response of the filter before tuning is shown in **figs. 11** and **12**. The calculated response which is given for comparison, was generated using **eq. 1** for each stage and then adding the four responses.

Normally, filter sections will need trimming for frequency and/or $Q$. In many low $Q$ filters ($\delta$ $0.3$), 5-per cent tolerance resistors will give quite satisfactory results without trimming. The only penalty may be slight center frequency error and perhaps a small amount of skew in the passband frequency response.

For the narrowband filters, and especially those with three or four stages, an audio generator, ac voltmeter, and frequency counter will help in trimming each stage independently to the required parameters. In the state-variable circuit, adjust both R3 values to set the center frequency, then use R5 to fix the $Q$. Remember $Q$ is the 3-dB bandwidth divided by $f_n$. For an MFB stage, adjust R3 to give the desired 3-dB bandwidth. Then adjust R2 to set $f_n$. Varying R2 has virtually no effect on the bandwidth, which means the $Q$ changes at the same rate as $f_n$. After tuning the RTTY demodulator input filter, the overall response was essentially indistinguishable from the calculated response.

## components

Generally, components should be the best you can get. Metal-film resistors and polystyrene or mylar capacitors are hard to beat, but may be *overkill.* Stay away from capacitors designed for bypass or coupling use; their tolerance is poor, as is their stability. Carbon resistors are usually adequate in all but the narrowest filters. For op amps, 741s are suitable (as are the 1458 dual versions and the quads like the

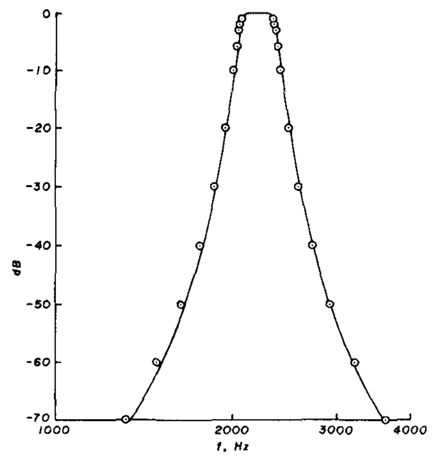


**fig. 11. Response as a function of frequency for an RTTY input filter. Solid line: calculated; dots: measured data before tuning.**

3303), when used within the limitations given above. The LM318 op amp gives much greater freedom from $Q$ drift (in the state-variable circuit) and $f_n$ drift (in the MFB circuit). Some of the new wideband fet-input op amps, such as the LF356, should be excellent performers. When external frequency compensation is required, use the values specified for unity gain amplifiers.

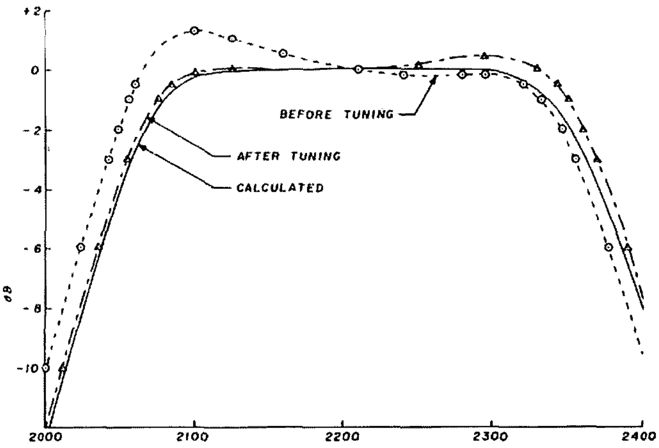


fig. 12. Passband response of an RTTY input filter.

When interfacing active filters, take care that the source impedance driving the filter is very low, i.e., less than 1 per cent of R1 in either circuit. Another op amp or a voltage follower provides an excellent driver. If the requirement for a low impedance can't be met, deduct the source resistance from the value of R1 in the first stage.

You've seen an easy-to-use method for designing stagger-tuned active filters to your own needs, and have learned to avoid some of the possible pitfalls. Now you can replace that filter you borrowed from someone else's circuit that never did work exactly the way you wanted.

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**ham radio**

# crystal-controlled phase-locked receiving converter

Circuit details and construction information for a converter that receives signals between 0-28 MHz when used with a receiver that tunes from 28 to 29 MHz

Over the years many amateurs have traded their old, general-coverage receivers for shiny new "ham-band-only" models. We've gained in stability, sensitivity, selectivity, dial accuracy, and many other attributes; but we've lost on frequency coverage. Except for a few narrow windows to the outside world, we can listen only to each other. Those with interests outside the amateur bands have had to use a second receiver (frequently an old general-coverage job) and put up with drift, bulk, and lack of accurate calibration. The *XPL Converter*, described here, is designed to work with a modern receiver to give the best of both worlds — extremely broad frequency coverage together with crystal stability and calibration accuracy. The converter receives all frequencies between 0 and 28 MHz when used with a receiver tuning 28 to 29 MHz. Construction is simple, straightforward, and inexpensive thanks to integrated circuits.

## description

The *XPL Converter* consists of a wide-range tuned input circuit, 60 kHz to 28 MHz; a local oscillator with injection frequency switch-selectable in 1-MHz steps from 29 to 56 MHz; and a mixer circuit with output 28 to 29 MHz feeding the receiver as a tunable i-f amplifier. A block diagram is shown in **fig. 1**. Local oscillator output is taken from a vfo (vco), which is phase locked to a 100-kHz reference oscillator through a counter chain preset by thumbwheel switches for band selection. The various sections are described in more detail later.

An example may help clarify the frequency conversion technique employed in the *XPL*. If the vco is set at, say, 38 MHz, the tunable i-f range of 28 to 29 MHz will allow reception of signals from 38-29 = 9 MHz to 38-28 = 10 MHz. A 9330-kHz signal in this range would be received at 38-9.33 = 28.67 MHz. The receiver tunes *backwards*, in that the low-frequency end of each range will be received at 29 MHz and the high end at 28 MHz. This turns out to be only a minor operating annoyance, however. Low-side injection could be used for forward tuning but only at the sacrifice of tuning range at the upper end.

The vco is phase locked to the reference crystal, so the local oscillator is of crystal quality as far as accuracy and stability are concerned. Any input frequency can be precisely located and will be stable within the accuracy and stability of the receiver on the 10-meter range. For most modern receivers, this means 1-2 kHz accuracy and a few hundred hertz drift on warmup. What a difference from the old general-coverage boat anchors!

## input circuitry

Input-circuit details are shown in **fig. 2**. A single-tuned circuit provides input selectivity for the *XPL*. Six switch positions cover 60-150 kHz, 150-450 kHz, 450-1400 kHz, 1.4-4.5 MHz, 4.5-10 MHz, and 10-30 MHz. The four high-frequency ranges use a commercially available coil set having high-impedance balanced antenna windings. On the two low-frequency ranges pi-section single-ended input circuits are used with rf chokes for the inductors. Tuning is by a miniature broadcast superhet variable capacitor having a total capacitance of about 560 pF with the two sections in parallel.

Two input traps are used: a balanced lowpass filter to eliminate TV/fm pickup and a series-resonant

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trap to eliminate overload from any one broadcast station. Additional suppression measures may be required in unusual situations.

There's no need to conform to the input circuit shown. In fact, the antenna tuning section from a scrapped general-coverage receiver could be used to

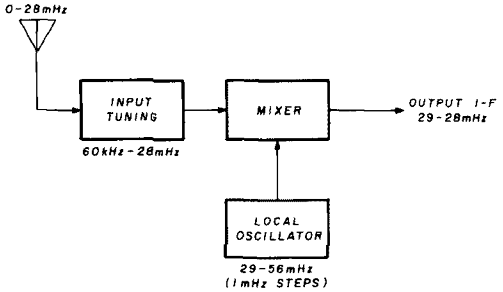



**fig. 1. Block diagram of the *XPL* converter.**

handle the high-frequency end of the range. The low end can be extended with larger inductors, but the tuning range for each band will be quite limited because of distributed capacitance in the coils. Two additional coils, however, will allow tuning to about 15 kHz.

## local-oscillator system

The heart of *XPL* is the local oscillator. This circuit consists of a voltage-controlled oscillator (vco), programmable divider chain, crystal-reference oscillator, and phase comparator. A block diagram is shown in **fig. 3.**

Phase-locked-loop operation has been well described in the literature, but a quick review may be worthwhile. A phase-locked loop is a feedback control system that measures the phase difference between two frequency sources and generates an error voltage that changes the frequency of one frequency source until the two sources are in phase synchronism. For continuing phase errors, the phase detector will function on frequency difference and steer the system into phase lock.

Two basic systems can be used to generate a selectable series of integrally related frequencies. If the phase comparator is sensitive to reference-oscillator harmonics, the controlled oscillator can be directly locked to a selected harmonic by first tuning it manually to a nearby frequency, then allowing the phase detector to lock up. This is the system used in several commercial receivers. The only objection from a construction point of view is that it requires a manually variable oscillator with dial calibration sufficient to resolve adjacent harmonics. A lock indication is also useful in identifying the proper harmonic.

A more direct way of generating the integrally related frequencies is to divide the controlled-oscillator frequency by programmable digital dividers before phase comparison to the reference frequency. If the oscillator frequency is divided by, say, 24 before the comparison is made, the effect is to lock the oscillator to the 24th harmonic of the reference frequency. In *XPL*, a fixed divide-by-ten and two programmable divide-by-n counters are used to enable lock from the 290th harmonic to the 560th harmonic of the 100-kHz reference frequency in steps of 1 MHz.

## oscillator and phase comparator

**Fig. 4** is the schematic for the reference oscillator and phase comparator. A 7400 quad NAND gate is used with a 100-kHz crystal to generate the reference frequency. There's no special merit to this scheme other than simplicity, and any convenient oscillator circuit could be used so long as it provides TTL output levels. In the circuit shown, the 0.0047 $\mu$F capacitor at the input to the last gate was necessary to eliminate a double-pulsed output to the phase comparator.

A Motorola MC4044P phase-lock chip was chosen because it offers TTL logic, a nonharmonic-sensitive

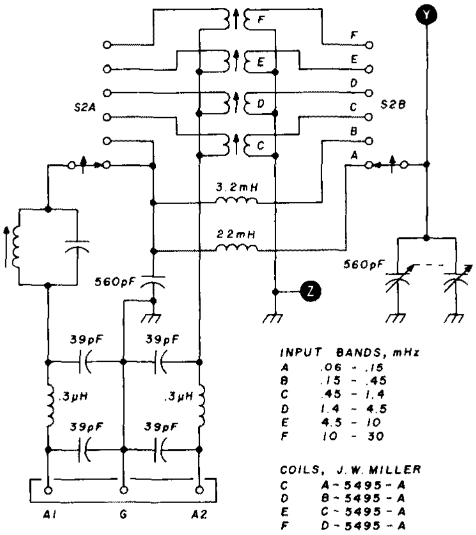



**fig. 2. Input-circuit schematic.**

comparator, and some internal auxiliary transistors. Also, its use in synthesizers has been described in recent articles.

Output from the MC4044P is buffered by an external 2N5457 fet follower and the internal emitter followers. The comparator has unity gain from the phase detector to the output. An active filter is backed up by two poles of rolloff for loop stability and high 100-kHz ripple attenuation. The reference

oscillator and phase comparator are supplied from an on-board regulator that provides both isolation and filtering.

## voltage-controlled oscillator

The voltage-controlled oscillator in the *XPL* (**fig. 5**) uses a Motorola MC1648L ECL chip designed for this service. Spectral purity requirements preclude a voltage-controlled multivibrator, so this chip was used with an external high-Q toroidal inductor and a Motorola MV1401 variable-capacitance diode or varicap. Since ECL has a very low logic swing, an

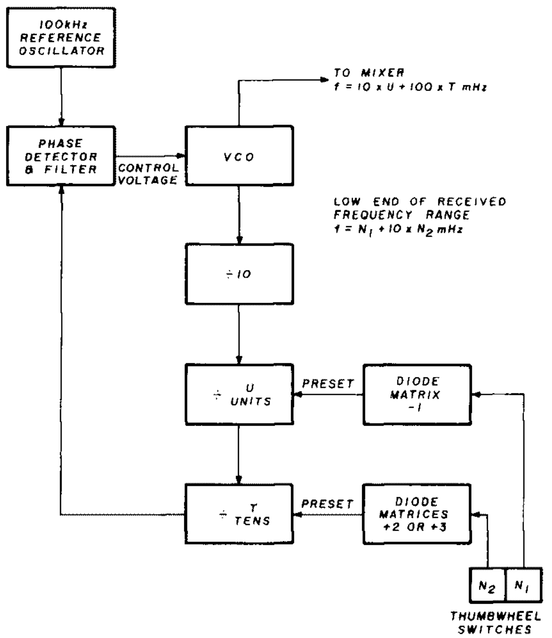



**fig. 3. Local-oscillator block diagram.**

output translator, 2N4403, is used to regenerate the TTL signal level. At this point you might ask whether the ECL chip is worth the effort. The answer is a qualified "Yes," since it functions from a two-terminal tank circuit and eliminates the need for fussing with feedback in a transistor oscillator.

The MV1401 varicap is rather expensive (in the $9.00 range), but it has a guaranteed 3:1 tuning range and high $Q$. This application requires only a 2:1 range, but allowances for temperature variation component tolerances, and other considerations make it necessary to have some overrange. Less-expensive limited-range diodes could be used, but they would require changing fixed capacitors to cover the tuning range for the vco. The inductor is a T-25 mix 6 toroid with four turns of no. 22-28 AWG (0.6-0.3mm) enameled wire.

An output to the mixer is taken directly from the 50 ohm vco output at pin 3. For the TTL counters, however, the swing is wrong. The sum of one diode drop and a base-emitter drop from the 5-volt-supply rail places the 2N4403 base voltage in the ECL logic voltage range. The 2N4403 collector voltage swings from 0.5V to about 3.5V to drive the counter. A 22-ohm base-emitter resistor aids junction recovery and cleans up the output waveform.

The entire vco section is quite susceptible to hum and modulation disturbances. For this reason, a separate voltage regulator is again used. The vco should be located well away from transformer fields or ac power wiring.

## programmable counters and translators

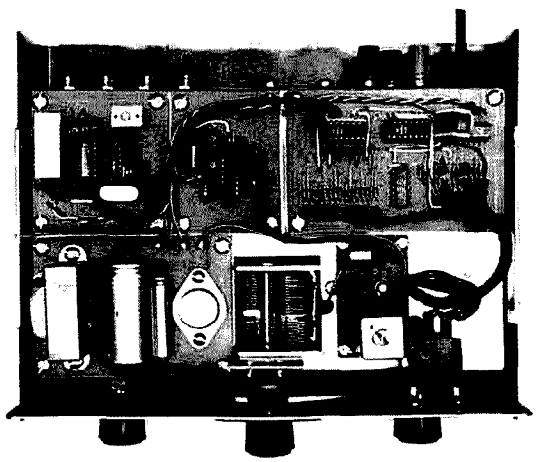
This circuit is shown in **fig. 6**. Before getting into counter details, a related matter must be considered. The count set into the preset counters must always be 29 (MHz) higher than the bottom end of the input tuning range, so that the switches can read input range directly. This requirement leads to the necessity of translating switch settings to the counters. **Table 1** summarizes the required relationships. If decimal switches are used, the offset of minus 1 in the units digit can be provided by simply rewiring into the decimal-to-BCD diode matrix, as shown in **fig. 6**. Note, for example, that a switch indication of 4 is translated to BCD $1+2=3$, which is 4 minus the required one unit. The 390-ohm resistors establish a TTL logic zero for open-switch positions.

The tens digit is somewhat more messy. Switch indications must be translated up by 3 *except* when the units position is zero, which requires an up-translation of only 2. Thus, 00 goes to 29, 01 goes to 30, 10 goes to 39, 11 goes to 40, and so on. Since a zero-units digit is translated to a 9 in the output, the presence of this 9 can be used to change the tens digit to an output lower by one integer.

Two sections of a 7400 quad NAND are used to accomplish this magic. A decimal-to-BCD diode matrix with an offset of +3 is used in conjunction with a second matrix with offset of +2. The proper matrix

**table 1. Relationship between switch settings and counters.**

| input range | switch readings ten | switch readings units | counter presets ten | counter presets units |
|---|---|---|---|---|
| 0-1 | 0 | 0 | 2 | 9 |
| 1-2 | 0 | 1 | 3 | 0 |
| 2-3 | 0 | 2 | 3 | 1 |
| 9-10 | 0 | 9 | 3 | 8 |
| 10-11 | 1 | 0 | 3 | 9 |
| 11-12 | 1 | 1 | 4 | 0 |
| 12-13 | 1 | 2 | 4 | 1 |
| 19-20 | 1 | 9 | 4 | 8 |
| 20-21 | 2 | 0 | 4 | 9 |
| 21-22 | 2 | 1 | 5 | 0 |
| 27-28 | 2 | 7 | 5 | 6 |



**Looking down on the *XPL* converter. Components and wiring are shown on top of the chassis.**

is chosen by clamping diodes from the two 7400 outputs. If a *units* 9 is present, the 9 bus is high and the 9 bus is low. Under this condition, the +3 matrix diodes are clamped low, and the +2 matrix diodes are released. For units digits other than 9, the situation is reversed.

Those familiar with counter techniques may immediately conclude that this is the long way around the barn, and so it is. A simpler solution to the translation would be to use a precounter set to 29 to delay activation of the programmable counters until the first 29 counts have passed. However, this requires two more counter chips, involves a clock gate, and is more difficult to troubleshoot if problems develop. The approach shown was devised with the less-experienced builder in mind, since troubleshooting of the diode matrices can be done with a vtvm.

Returning now to the counters, a high-speed 74196/8290 is used as a divide-by-10 prescaler to get the signal into TTL frequency range. This unit has a typical toggle frequency of 75 MHz and handles the 56-MHz maximum input with little effort. The programmable counters, 74192s, are preset by their respective diode matrices. Both are operated in the countdown mode and are cascaded and loaded through the borrow outputs.

Counter operation is as follows: The $C_0$ output of the 74196/8290, pin 2, goes high once every ten input pulses from the vco. Each output pulse causes the units 74192 to count down by one count. When the count reaches zero, the borrow output goes low between pulses and causes the tens 74192 to count down by one count. It, too, generates a borrow pulse after reaching zero, and it is this pulse that's used to reset the system. The borrow pulse from the tens counter is used to load the preset number into each counter.

As an example of operation, suppose the thumbwheel switches are set to 08 (8-9 MHz range). The units counter will be preset to a count of 7 (8-1) and the tens counter to 3 (0+3). Following a reset pulse, the units counter will count down one count every ten vco cycles and first generate a borrow pulse after 70 vco cycles.

After this first 70 cycles from the reset pulse, the counter will again generate a borrow pulse every 100 vco cycles. Each of these borrow pulses causes the tens counter to count down by one count from its preset of 3 and to generate its own borrow pulse after 70+100+100+100 vco cycles. The tens counter borrow and reset pulse thus occurs 370 vco cycles *after* the first reset and then immediately resets the counters again.

The tens counter borrow output frequency is equal

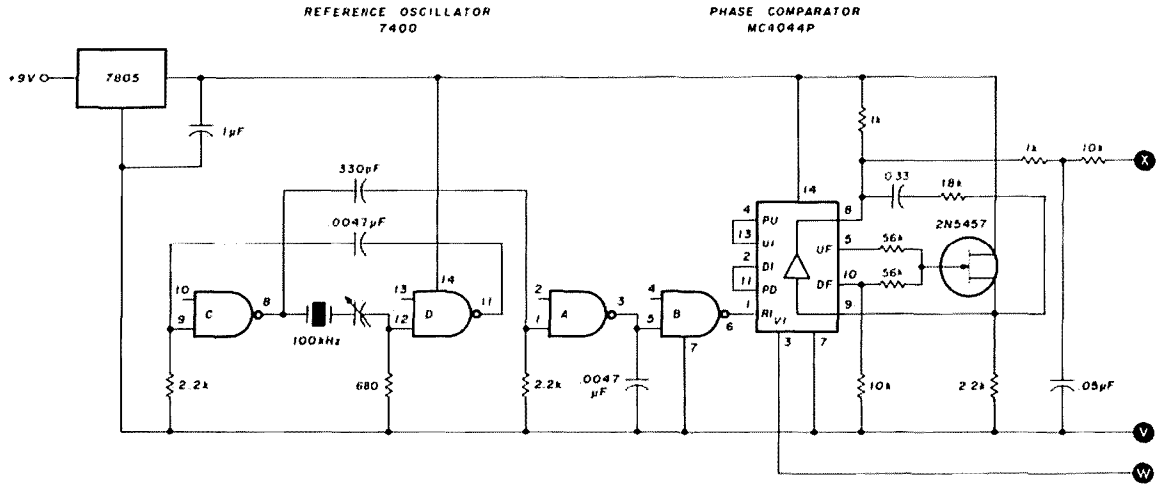



**fig. 4. Reference oscillator and phase-comparator schematic.**

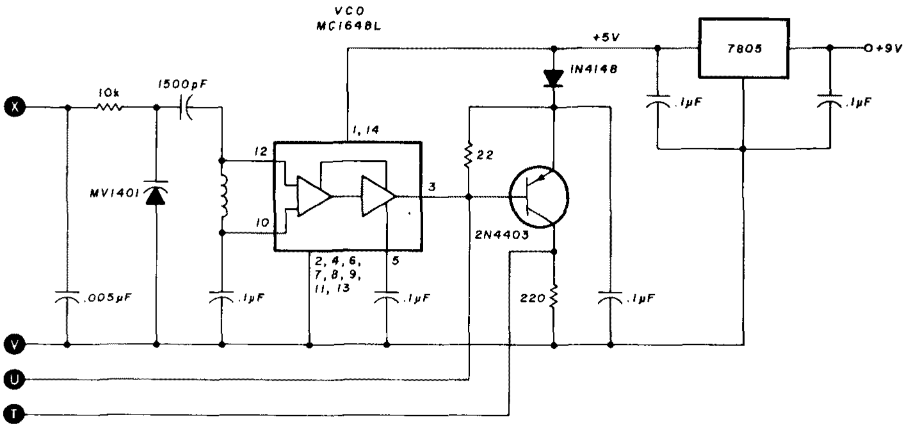



fig. 5. Voltage-controlled oscillator schematic.

to the vco frequency divided by (10xU + 100xT), where U and T are the units and tens presets respectively (7 and 3 in our example).

The tens-counter borrow output is also used to feed the phase comparator so that the vco frequency is locked to 10x7 + 100x3 = 370 times the reference frequency of 100 kHz. The vco is thus locked at 37 MHz, which is the required local-oscillator injection frequency for receiving 8 MHz with a 29-MHz i-f.

Resistor values shown for the diode matrices are fairly critical. Germanium diodes would provide more margin, but the circuit works well as shown. If the same nominal values are used, no problems should be experienced. Power for the counters and translators is provided by still another 5-volt regulator. This system draws several hundred milliamperes and may need a regulator heat sink.

## mixer

A dual-gate, diode-protected mosfet is used for the mixer (**fig. 7**). The 40673 has good intermodulation characteristics and is simple to use. Output from the drain is taken through an output transformer, broadly resonant at 28.5 MHz, which provides a low impedance output to the receiver. This stage is powered directly from the 9-volt power supply, since decoupling is not a problem. An output switch pole allows the receiver input to be connected to the converter or to a high-frequency antenna. A second pole is used to ground the high-frequency antenna to minimize pickup when the *XPL Converter* is in use.

## power supply

All operating power for the *XPL* is derived from a 12-volt transformer and bridge rectifier at about 10 volts (**fig. 8**). A 2N3055 is used as an active filter to reduce ripple. This transistor is much larger than required, but it's cheap, readily available, and needs no heatsink. If the 9-volt rail is not reasonably clean, the received signal may be hum modulated. Ac input is switched by a third pole of the IN-OUT switch, and a pair of 0.02 $\mu$F capacitors are used for line bypassing. Note that these capacitors should have 600-volt ratings.

## construction

Each circuit section was built on a separate printed-circuit board for easy testing and debugging. There's no real need to do this, however, and a single PC board might be easier to handle mechanically. The layout shown is also more compact than necessary. The entire unit could be built on perf board if generous ground conductors are used.

Coax cable should not be used for interconnecting circuits except for the vco output to the mixer, mixer output to the receiver, and input from the hf antenna. Other leads should be run in twisted pairs of no. 22-26 (0.6-0.4mm) hookup wire to minimize shunt capacitive loading on the TTL gates and to reduce inductive pickup in the phase-comparator circuitry. Coax cable should not be used in the input circuit, since the high capacitance of this cable could appreciably decrease the tuning range.

The 28.5-MHz output coil was a junk-box relic of unknown parentage. Any coil with a turns ratio of about 5:1 with a slug capable of resonating at 28.5 MHz will do. An inductance of about 3$\mu$H is required.

Most of the parts for the *XPL* are available from surplus houses or other *ham radio* and *QST* advertisers. The MC4044P, MC1648L, and MV1401, however, will probably have to be ordered from a franchised Motorola distributor. Total cost is about $20 for these items.

The individual regulators were Motorola types, but various National LM-series are equally satisfactory and widely available. The 2N4403 transistor can be

replaced with almost any high-frequency pnp transistor. Similarly, the 2N5457 can be replaced by other N-channel jfet devices, such as the MPF102 series.

All signal diodes should be 1N4148/1N914 or similar silicon computer *diodes*. As mentioned earlier, germanium diodes can be used in the diode matrices if desired. Power diodes are low voltage, plastic-lead-mounted types.

Capacitors can be ceramic units except for the antenna input capacitor (560 pF) and the vco 1500-pF capacitors which should be of low-loss polystyrene or mica construction.

The entire counter and phase-lock system could be designed around CMOS circuitry except for the vco and the prescaler. CMOS chips are not widely available in surplus outlets but are rather inexpensive when purchased new. Power supply current could be considerably reduced by shifting to CMOS, and the diode translators could be run at a much higher impedance level.

## adjustments and troubleshooting

The power supply forms a logical first item if the *XPL* is built in steps. Output voltage must be at least

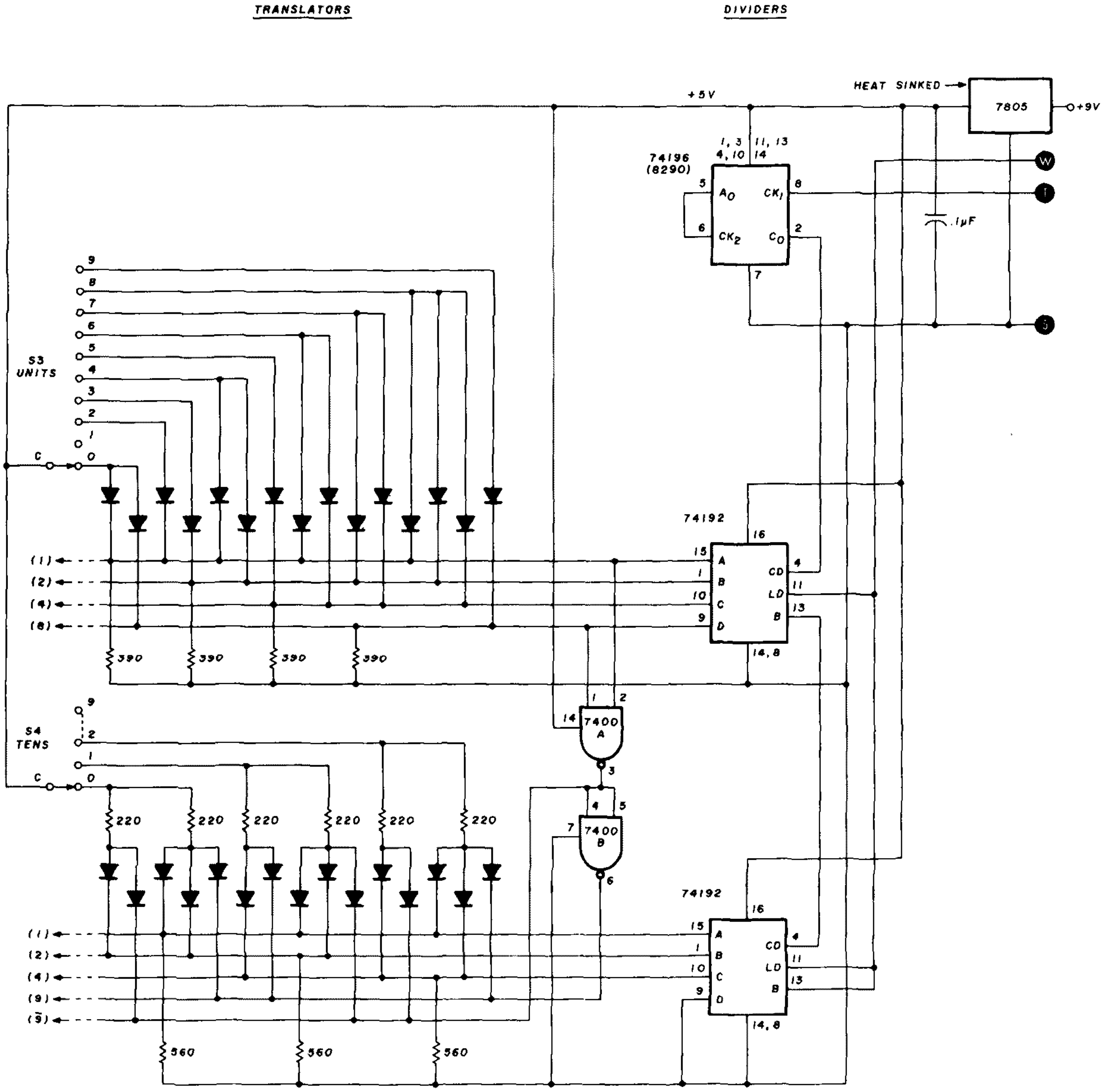



**fig. 6. Schematic of the *XPL* programmable counters and translators.**

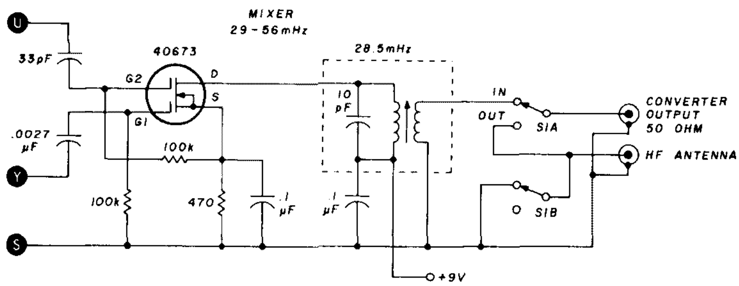



fig. 7. Mixer schematic.

7 volts to allow the individual regulators their required 2-volt input margin over 5-volt regulated output voltage. Frequency calibration of the reference oscillator can be done by zero beating with WWV or with WCFL, Chicago, on 1000 kHz. The vco should be checked for range by coupling a grid-dip oscillator to the toroidal coil. An input of 0.5 to 3.5-volts positive, derived from a separate source, should drive the vco from 25 to 60 MHz or so. If the frequency range is off, toroid turns may be trimmed or the 1500-pF capacitor value changed to suit. Input-coil slugs should be adjusted to allow coverage of all input frequencies with a bit of overlap.

Operation of the translators can be checked with a vtvm. Logic zero must be 0.8 volt or less, and logic 1 must be 2.4 volts or more — standard TTL levels. Preset counter operation can be checked with a triggered oscilloscope and a low capacitance (10X) probe. The 74196 output pulses will be visible on most inexpensive scopes.

## operation

For best results, a good antenna system should be used with the *XPL*. One of the best is an 80-meter inverted V or dipole with open-wire feeders. Except for those few frequencies at which the antenna happens to be an odd number of quarter wavelengths long, its impedance will be quite high. Thus, the high capacitance of a grounded coax antenna feeder would result in serious signal attenuation. A simple long-wire antenna can be used if the end is brought directly to the *XPL* input terminals. If a separate antenna system isn't available, any ungrounded antenna feeder system can be used by connecting either lead to the **A1** antenna terminal. Terminal **A2** should be grounded for single ended inputs.

The input circuit should be calibrated at least roughly so you can be sure the desired signal is being peaked. The input circuit provides the only rejection for 10-meter signals present on the antenna. Above about 15 MHz, this rejection may be inadequate to prevent strong 10-meter signals from coming directly through the mixer to the i-f. A resonant trap or a loosely coupled input circuit can be added if this problem proves troublesome. A balanced mixer would reduce the feedthrough, but the added complication seemed unnecessary. I suggest this as an alternative approach for those interested in experimentation.

## final remarks

The *XPL Converter* has been fun to use. Broadcast stations pop up exactly where they are supposed to be. The *Selected Cities Weather Summary*, broadcast from Miami on RTTY has been interesting to print and peruse. WWV is available on all frequencies for calibration or a check on propagation conditions.

Aviation weather and general information is broadcast on the local low-frequency range station. Every international shortwave band can be received. International air-route traffic control from Miami and New York can also be monitored. And, near the top end, you can even listen to CB operations.

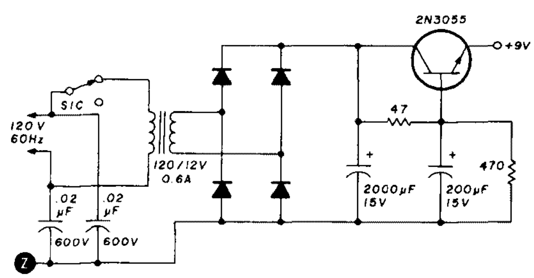



fig. 8. Power-supply schematic.

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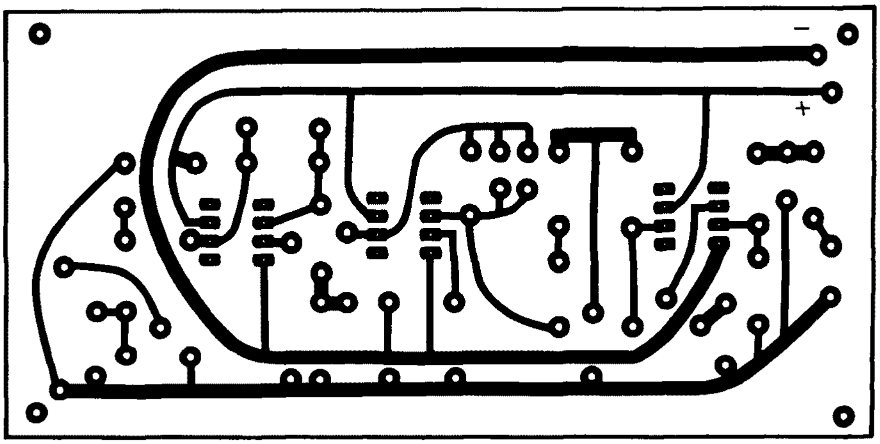
**ham radio**

# short circuit

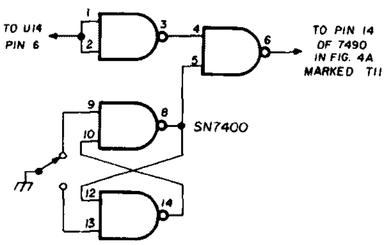



## RTTY time/date printout

An important point was missed in **table 1** of the RTTY printout article which appeared in June, 1976, *ham radio*. Pin 4 of U15 should not be grounded but should have the appropriate BCD information for the tens of minutes digit.

As shown, the ten, minutes digit will only display up to 39 minutes instead of 59 minutes. In **fig. 4A**, pins 6 and 7 of the 7490s must be grounded; otherwise the circuit will only print the 19th as the date.

Advancing the date by moving the clock is a very tedious process. Overshooting will mean doing the entire thirty days over again. The diagram below shows a circuit that will permit




you to advance the date by one day with the flip of the switch. When the date resets at the end of the month, flipping the switch will advance the clock to Ø1, much easier than advancing the digital clock a complete 24-hour period. Note that this advance circuit is designed to work with a low input so the date advance must be done before 2000 hours.

## pi network design and analysis

**Eq. 8** in the pi network design article, September, 1977, *ham radio*, should not have the radical sign on the right hand side of the expression; it should read

$$R1 \text{ (at minimum point of } X_{C1} \text{ curve)} = R_{1B} = \frac{2X_L^2}{R_2} \qquad (8)$$

Also, **eq. 12** should read as follows:

$$X_L = (R1+R2)\ \frac{Q_o^2 + \sqrt{Q_o^2 - (Q_o^2+4)\left(\frac{R2-R1}{R1+R2}\right)^2}}{Q_o^2+4} \qquad (12)$$

## direct output synthesizer for two meters

In **fig. 4** of the direct output synthesizer in August, 1977, *ham radio*, the lines connected to pins 10 and 8 of U3C have been transposed. For correct operation, pin 10 is connected to the pin 9s of the 74161s, and pin 8 of U3C is connected to pin 1 of U2B. On U1, pin 2 is the input from U3D and pin 1 should be connected to the junction of the 100 and 360 ohm resistors. U1 may exhibit some temperature and voltage sensitivity at times causing the divide-by-21 function to become a divide-by-22. This problem can be cured by either of two methods: putting a 330 pF capacitor from pin 2 of U1 to ground or replacing U3 with a 74L00 instead of the 7400. U8 is a 7483, not a 7473. In **fig. 6**, the 0.1 μF capacitor connected to pin 2 of U18 should be a 0.01 μF disc capacitor. Also, the 40k-ohm resistor on the output of U18B should be 10k.

## serial converter for 8-level teleprinters

The serial converter in August, 1977, *ham radio*, uses a 74121 for U16, not a 7474.

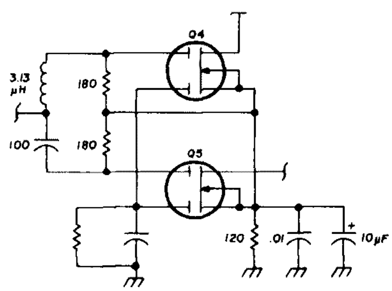
## audio frequency speech processing

The circuit board layout for the audio speech processor in August, 1977, *ham radio* was missing several connections. The diagram above shows the correct circuit board layout. The output is taken from the center of R13 and not as shown in **fig. 5** in the article. The numbering for the pins of the ICs in the schematic diagram should be changed to correspond with the 8-pin mini DIPs used on the finished board.

| fig. 3 | change to |
|---|---|
| 5 | 3 |
| 4 | 2 |
| 6 | 4 |
| 10 | 6 |
| 11 | 7 |

## phasing-type single-signal detector

In **fig. 2**, page 72 of October, 1976, *ham radio*, the two 180-ohm resistors should be connected between gate 2 and the source of the dual-gate mosfet as shown below. Also, gate number 1 is not connected to the source.



### spectrum analyzer

There are several errors in the spectrum analyzer construction article which appeared in the June, 1977, issue. The 75.1 ohm resistor in the rf attenuator should be 71.5 ohms; the six 69.1 ohm resistors should be 61.9 ohms (**fig. 10**). The i-f attenuator should have three, *not two*, 20 dB sections (like the rf attenuator).

The mixer diodes used by the author are Hewlett-Packard part number 5082-2900; most any hot-carrier diodes should work if they are all the same type.

The crystal in the second local oscillator is 150 MHz ±2 MHz; the crystal in the third local oscillator is 39.3 ±1 MHz. The 10k resistor associated with CR401 should go to switch S601A, the 250 kHz position; the same for the 10k resistor associated with the second crystal filter, Y401 (**fig. 11**). The 2.4k resistor in series with CR402 should go to switch S601A, the 10 kHz position. The coil located near CR403, and the switch contacts near R402, are parts of the same relay.

Large size Xerox copies of the top and bottom chassis photographs are available from *ham radio*, and will be sent to interested readers upon receipt of a self-addressed, stamped envelope.

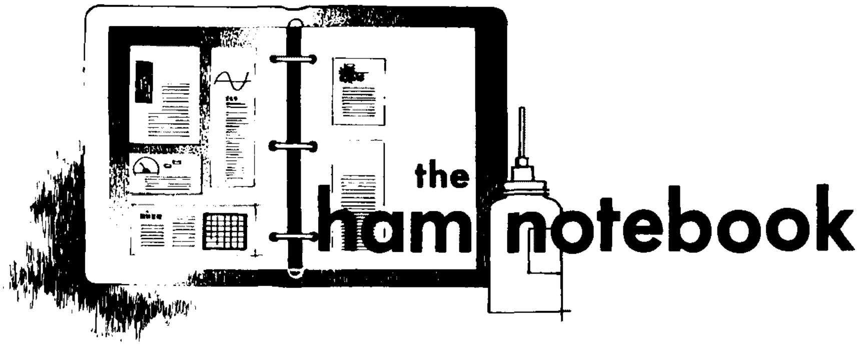
### reducing IMD in high-frequency receivers

The 3-dB pad between the local oscillator input and the balanced mixer, in **fig. 6** on page 30 of the March, 1977, issue of *ham radio*, should have the values transposed (the series resistor should be 18 ohms, the shunt resistor 300 ohms.)

### bandspreading techniques for resonant circuits

In **eq. 19** on page 49 of the February, 1977, issue of *ham radio*, the term $C_r$ should not be included under the radical sign. The equation should read:

$$C_p = \frac{\sqrt{C_q + C_r^2}}{2V} - C_r$$




## simple formula for microstrip impedance

In many amateur vhf and uhf applications strip transmission lines etched on printed-circuit board are used for impedance matching and as components in tuned resonant circuits. Although several methods are available for calculating the characteristic impedance of microstrip transmission line, the formula derived by Sobol[1] is the most popular. It has been widely publicized in Motorola Semiconductor's application notes and appeared recently in *QST*[2]. Sobol's equation:

$$Z_o = \frac{120\pi h}{\sqrt{\epsilon_r}\, w(1 + 1.735\epsilon_r^{-0.0724}\ w/h^{-0.836})}$$

where $w$ is strip width, $h$ is the dielectric thickness, and $\epsilon_r$ is the relative dielectric constant of the substrate.

Sobol's equation gives $Z_o$ as a function of microstrip geometry, but in practical applications you usually need to know what size microstrip is required for a given impedance. Since the equation can't be solved directly for $w/h$, an interactive trial-and-error solution is necessary. This can be done rather quickly with a high-speed computer, but an iterative solution with a programmable calculator such as the HP-25 may require a minute or more — an iterative solution with a non-programmable calculator is impractical.

Some time ago N6TX (ex WA6UAM) sent me an iterative HP-25 program for Sobol's microstrip equation which provided acceptable accuracy for most design work. This program begins at $w/h = 1$ and iterates out to the required value. Therefore, for low and high values of $Z_o$ a solution requires considerable calculation time. To reduce calculation time I decided to see if I could develop a simple equation for an approximate value of $w/h$ which could then be refined with the HP-25 program. My first step was to rewrite Sobol's equation as

$$Z_o = \frac{120\pi}{\sqrt{\epsilon_r}\frac{w}{h}}\left[\frac{1}{1 + 1.735\,\epsilon_r^{-0.0724}\ w/h^{-0.836}}\right]$$

By inspection, to a first approximation $Z_o$ is equal to the first term on the right-hand side of the equal sign; the term inside the parenthesis is a modification term which is a function of both $\epsilon_r$ and $w/h$. Designating the quantity $(1.735\epsilon_r^{-0.724}w/h^{-0.836})$ as $K$, eq. 1 was rewritten as

$$\frac{120\pi}{Z_o\sqrt{\epsilon_r}} = \frac{w}{h}\,(1 + K) = \frac{w}{h} + K \cdot \frac{w}{h}$$

All that remained was to find a value for $K \cdot w/h$ which satisfied varying values of $\epsilon_r$ and $w/h$. After calculating several tables of values, it was ap-

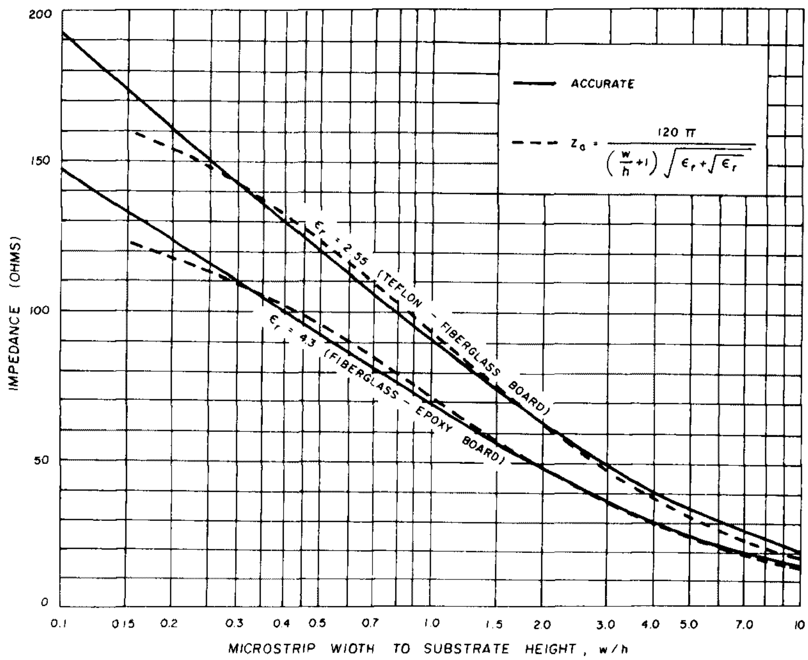



**fig. 1. Microstrip impedance calculated with simple formulas developed by W1HR (dashed lines), as compared to actual impedance (solid line). For $\epsilon_r > 4$, accuracy is very good for w/h > 0.2.**

1. H. Sobol, "Extending IC Technology to Microwave Equipment," *Electronics*, March 20, 1967, page 112.
2. R. Olsen, N6NR, "Designing Solid-State RF Power Circuits," *QST*, September, 1977, page 15.

parent that $K \cdot w/h = 1$ would give the desired results. Substituting and rearranging terms yielded the expression

$$\frac{w}{h} \approx \frac{120\pi}{Z_o \sqrt{\epsilon_r}} - 1 \qquad (3)$$

When this equation was plotted on graph paper and compared to a graph of Sobol's equation, the similarity was much closer than I expected — the curve had essentially the correct shape, but all values were slightly larger than those given by Sobol's formula. This was the desired result; rewriting the HP-25 program around eq. **3** considerably reduced calculation time.

Later it occurred to me that it might be possible to further factor **eq. 3** to obtain a more accurate formula for microstrip impedance. After calculating numerous tables of $Z_o$ vs $w/h$ and $\epsilon_r$, and inspecting the values, I found that the impedance of microstrip etched on a substrate with $\epsilon_r > 4.0$ could be approximated within a few per cent by the following equations:

$$\frac{w}{h} \approx \frac{120\pi}{Z_o \sqrt{\epsilon_r + \sqrt{\epsilon_r}}} - 1 \qquad (4)$$

$$Z_o \approx \frac{120\pi}{(\frac{w}{h} + 1)\sqrt{\epsilon_r + \sqrt{\epsilon_r}}} \qquad (5)$$

For microstrip etched on glass-epoxy circuit board ($\epsilon_r = 4.8$), these equations can be reduced to

$$\frac{w}{h} \approx \frac{142.6}{Z_o} - 1 \qquad Z_o \approx \frac{142.6}{\frac{w}{h} + 1}$$

For Teflon-fiberglass circuit board ($\epsilon_r = 2.55$) the simplified expressions are

$$\frac{w}{h} \approx \frac{185.1}{Z_o} - 1 \qquad Z_o \approx \frac{185.1}{\frac{w}{h} + 1}$$

The dielectric constant of Teflon-fiberglass is below the value recommended for these equations, but accuracy is still acceptable for many applications.

These formulas can be solved quickly by hand (or with a simple four-function calculator), and should be a big help to amateurs who want to design their own microstrip circuits. They can also be used to determine the approximate impedance of circuit traces for digital logic boards (for best results the $V_{CC}$ and ground lines for TTL should have low impedance).

The accuracy of these simplified equations is surprisingly good. As shown in fig. **1**, for $w/h > 0.2$, the simplified formulas are within a few per cent of the impedance calculated with more accurate equations; this covers the microstrip impedance range most commonly used in radio communications work. With fiberglass-epoxy board the formulas are within about 1 ohm of the exact expression for all values of $Z_o$ below 60 ohms. The values for Teflon-fiberglass board are somewhat less accurate, but are still acceptable for most amateur work.

**James R. Fisk, W1HR**

## improved (vfo) stability for the Atlas 180

Early versions of the Atlas 180 transceiver have exhibited poor vfo stability with a varying dc supply voltage. In some cases, the vfo will actually be frequency modulated at dc input voltages below **13** volts. Atlas owners can check for this condition by listening to a signal or the calibrator beat note and adjusting the dc supply from about 11.5 volts to 14.5 volts. A 500-milliampere supply is more than ample to operate the receiver. The test can also be made in the car by first setting up the beat note with the engine off and then starting the engine. After a few moments the battery system will come up to full-charge voltage of 14.5 volts. Any change in pitch during this time indicates poor vfo power supply regulation. The units in which this is most likely to occur are those which use a 10-volt regulator circuit consisting of a transistor with a 10-volt Zener on the base.

The solution to the problem is to remove the 27-ohm decoupling resistor (R401 in my Atlas 180) on the vfo board (PC-400), and replace it with a 78L08ACP low-power 8-volt regulator. The wire that previously connected to the 10-volt bus is then reconnected to the 13-volt bus. After making this change, retuning is unnecessary for dc inputs of 11.5 volts to 14.5 volts, and there are no reports of frequency modulation when operating mobile without the engine running. There is no other noticeable change in the operation of the vfo due to the 8-volt rather than 10-volt supply.

**Dave Sargent, K6KLO**

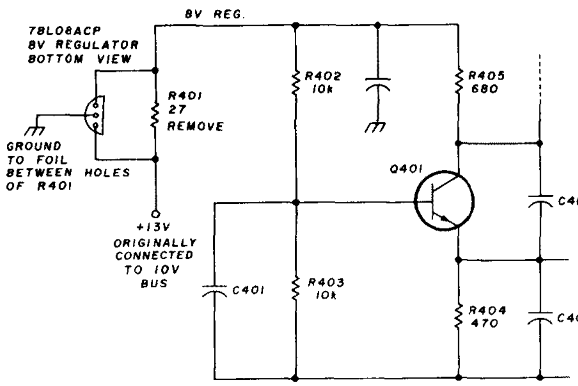



**fig. 1. Modification to the Atlas 180 Vfo power supply to prevent any frequency modulation due to voltage changes. The 78L08ACP voltage regulator is used to prevent voltage changes. It is fed from the normal 12.6-volt dc supply.**

# NEW products

For literature on any of the new products, use our *Check-Off* service on page 150.

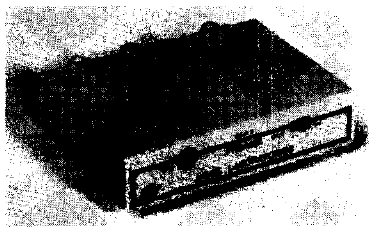
## crystal filters

Sherwood Engineering has announced two new additions to their crystal filter line. As complements to the CF-600/6, the new CF-2.6K/8 or CF-2.3K/8 crystal filter sets will replace the normal 8-kHz wide first i-f filter in the Drake R-4C. Each set has two filters, USB and LSB, that must be switched for the correct sideband. The individual filters are 8-pole crystal-ladder filters.

The CF-2.6K/8 is a set of ssb-bandwidth filters that are approximately 200 Hz wider than the normal second i-f phone filter. This allows a limited amount of passband tuning, while still reducing the second i-f bandwidth from 32 kHz, at −60 dB, to approximately 4 kHz. The other phone filter pair is the CF-2.3K/8, which is slightly narrower (100 Hz nominally) than the second i-f filter. Having the new filter sharper than the normal filter produces the equivalent of a 2 to 2.1 kHz filter, with 16 poles distributed over two frequencies. The passband tuning is then used to align the center frequencies, of the two filters, for proper cascading. This narrow combination offers the ultimate in phone selectivity. The bandwidth using the CF-2.6K/8, with the normal phone filter, is 2.3 kHz, at −6 dB, and 3.1 kHz at −60 dB; the bandwidth for the CF-2.3K/8 is 2.1 kHz and 2.9 kHz, at the 6 and 60 dB points. The additional advantages gained by distributing selectivity over two i-f frequencies are: virtual elimination of the chance of overloading the second mixer, and elimination of off-frequency signals that leak around the normal second i-f filter.

In addition to offering the basic filters, Sherwood Engineering also sells switching kits for the first i-f filters. The simplest arrangement is for the operator who wants to switch only between the two ssb bandwidth filters (CF-2.3K/8 or CF-2.6K/8). Custom-designed kits are also available to permit switching of all first i-f filters, 8 kHz, 2.6/2.3 kHz, or 600 Hz. Prices for the new filters are $120. The basic switching kit is $29.00 with the cost increasing approximately $25.00 per additional filter switched. Exact price quotes are given based on an individual's needs. For more information, contact Sherwood Engineering, Incorporated, 1268 South Ogden Street, Denver, Colorado 80210.

## two-meter preamplifier



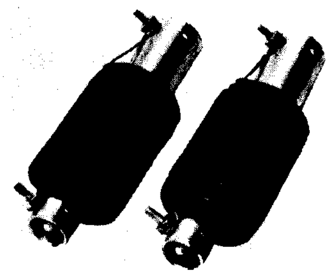
A new two-meter preamp has been introduced by Janel Labs. This preamp is specially designed to improve the sensitivity of transceivers and includes bypass circuitry for carrying transmit power through the unit. The preamp has a low noise figure, which gives excellent sensitivity for weak signals. An adjustable delay circuit (similar to that used in VOX circuits) allows for its use on all modes — f-m, ssb, am and CW.

The gain of the QSA 5 has been optimized for transceivers. It has a 15-dB gain level, which is sufficient to improve the sensitivity as much as practical but low enough to avoid creating overload problems.

A front-panel switch on the QSA 5 disables the preamp from the antenna line. This switch allows you to reduce gain on local signals and also allows experimentation on weak signals. A LED pilot light indicates when the preamp is in the line. This same LED also indicates when transmit power is being sensed.

The QSA 5 preamp is available from Janel Laboratories, 3312 S.E. Van Buren Blvd., Corvallis, Oregon 97330. The QSA 5 is available from stock at $39.95 plus postage. A full one-year warranty is provided. Specifications are available upon request.

## multiband antenna coils (40 through 10 meters)



Microwave Filter Company announces a set of antenna coils that will convert an amateur antenna from a single-frequency band of limited operation to operation on all amateur hf bands (40-10 meters).

Known as Reyco antenna coils, they are designed to shorten the overall physical length of an original single-frequency-band antenna. Model numbers are KW-40, 20, 15, and 10. Used in pairs, the model KW-40 coils will give flexibility of operation on all five hf amateur bands. Ideal performance is obtained by using all four coil pairs (KW-40 through KW-10).

In today's crowded apartment and suburban communities, the shortened antenna using Reyco multiband coils provides flexibility in minimum space. For additional information, write Microwave Filter Company, 6743 Kinne Street, East Syracuse, New York 13057.

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# semiconductors



# single sideband

# television



# transmitters and power amplifiers

## general

## high-frequency transmitters



## vhf and uhf transmitters



## transmitters and power amplifiers, test and troubleshooting



# troubleshooting

# vhf and microwave

## general



# vhf and microwave antennas



# vhf and microwave receivers and converters

## vhf and microwave transmitters